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ELECTRICALLY-SMALL, SUPERCONDUCTING ANTENNAS

Bernhard M. Schmidt

Dayton Electronic Products Company, Inc.
117 E. Helena Street
Dayton, Ohio 45404

Contract No. AF 19(628)-5893

Project No. 4600
Task No. 460010

FINAL REPORT

Period covered: 1 March 1966 - 28 February 1967

April, 1967

Contract Monitor: Charles E. Ellis

Distribution of this document is unlimited

Prepared

for

AIR FORCE CAMBRIDGE RESEARCH LABORATORIES
OFFICE OF AEROSPACE RESEARCH
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FOREWORD

This work was performed for Air Force Cambridge Research Laboratories, Office of Aerospace Research, United States Air Force, Laurence G. Hanscom Field, Bedford, Massachusetts, under Contract No. AF 19(628)-5893, Project No. 4600, Contract Monitor: Charles E. Ellis, CRDM, Electronic Systems Division (ESKK), by Dayton Electronic Products Company, Inc., 117 E. Helena Street, Dayton, Ohio 45404.

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Mr. George H. Kramer of Dayton Electronic Products Company, Inc., organized and conducted all experimental research involved in the reported work. Mr. James D. Hoeffel of Dayton Electronic Products Company, Inc., assisted in the theoretical aspects of the program. Their significant assistance is gratefully acknowledged.

ABSTRACT

The advantages and limitations of electrically-small, superconducting antennas have been investigated. The study led to a consideration of miniaturization, physical shape factors, long range magnetic coupling, maximum signal levels, antenna-receiver interface problems, materials, structures, and potential antenna applications of the quantum effects in superconductors. In addition, natural cooling and superdirectivity were incidental but relevant topics. In general, it was found that the possibility for miniaturization represents the principal advantage of the superconducting antenna, especially at the lower frequencies where antennas often are electrically-small through physical necessity. Radiation efficiency is increased in transmitting antennas, but at the expense of bandwidth. The degree of usefulness of superconductivity in receiving antennas depends considerably on the low noise properties and input impedance of the receiver and on the environment of the antenna. Any cooling improves the performance of the antenna; in fact, except in the case of the perfect receiver, there is some question as to which is more effective, the low temperature or the superconducting mode.

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SECTION I

SUPERCONDUCTIVITY AND ANTENNAS

1.1 Introduction

The motivation for this study of electrically-small, superconducting antennas was a conviction that superconductivity could alleviate some of the technological difficulties which prevent the practical realization of certain attractive theoretical properties of electrically-small antennas. Of primary interest at the start was the physical size-independent, effective cross section of the incremental dipole. It is well known that the aperture of such an element is approximately $0.12\ \lambda^2$ regardless of its actual geometrical cross section or size.

The aperture of the much larger half-wave dipole is only slightly greater in value ($0.13\ \lambda^2$). When one considers that the respective radiation patterns also are nearly the same, it would seem logical that efficient miniaturization could be accomplished by using the incremental antenna. Under conditions of maximum power transfer this potential advantage is lost because the radiation resistance of the incremental element is considerably smaller than its ohmic resistance; accordingly, the application of superconductivity comes to mind because of the vanishing ohmic resistance of a superconductor.

Another phenomenon of interest is the behavior of a high-Q tuned circuit. The concept of aperture or cross section can be applied to a simple tuned circuit. If the Q is low, say approximately 10, the circuit is relatively insensitive to its surroundings. Other circuits or metallic objects can be located nearby without, it seems, greatly disturbing the tuned circuit. For example, multistage, low gain, relatively wide bandwidth tuned amplifiers can be constructed without elaborate interstage shielding. Quite often, exposed coils may be used; the major feedback problem is associated with coupling impedances in the common power supply circuit. Now, when the Q is high, say 200 or

greater, the behavior changes drastically. The circle of influence is much enlarged and carefully designed shielding is required to prevent interactions.

Closer analysis shows that the LC circuit is simply a tuned incremental magnetic dipole to which the opening statements also apply; however, there are interesting areas of interpretation that could be explored. In circuit analysis we deal with interactions between coupled circuits via the coefficient of coupling concept. We observe that only a small coefficient of coupling is required for efficient coupling between two high-Q tuned circuits. When magnetic coupling is employed the degree of it can be controlled by physical separation of the two circuits, and it is here that the interesting problem arises. Ordinarily, the Q-factors of the circuits are low and only small physical separations are involved. However, extremely high Q-factors permit unusually large physical separations and the time retardation effects cannot be overlooked, as they usually are.

Superconductivity provides a means for realizing stable extremely high-Q tuned circuits; consequently, the interpretation of time retardation effects becomes a significant problem; it extends beyond one of mere academic interest. It is true, of course, that there are other ways of achieving extremely high-Q tuned circuits. Active elements, positive feedback circuits, and the like can, in principle, permit unlimited Q-factors; however, these are not stable circuits and the dynamic range decreases rapidly with increasing effective Q until finally the achievable Q-factor is noise limited. This is an important difference that gives the superconducting circuit a unique advantage in a number of potential applications that will be discussed in later sections. It is the possibility for large physical separation that makes the problem relevant to the study of electrically-small, superconducting antennas.

It is not likely that the antenna designer will ever put aside thoughts about that elusive goal; the superdirective antenna. The hard-headed, unimaginative engineer or scientist continues to tell us that this theoretically predictable concept is practically unrealizable; however, it is contrary to the human spirit to ignore a challenge of nature. Speculation will continue in spite of the fact that only small gains have been made after years of effort. The superdirective antenna enters our area of concern because superconductivity may be capable of making a contribution toward its

practical realization. By permitting a reduction of physical size, superconductivity makes possible extremely close spacing of elements in a superdirective array. This may warrant new experimental efforts in that direction.

Atomic and nuclear radiators and absorbers are of particular interest in this study of electrically-small, superconducting antennas because they can be looked upon as natural electrically-small antennas if we generalize our concept of "antenna". Normally one associates optical radiation with such entities, but the transitions involved in radiation and absorption are by no means limited to such high frequencies. As we shall see, low radio frequencies can be absorbed and radiated by single atoms and by single nuclei. These represent the ultimate in miniaturization and there may be much to learn by studying such natural processes, and especially so because the persistent superconducting state bears some resemblance to the persistent atomic state.

In subsequent sections we shall discuss what are called "conventional" and "non-conventional" applications of superconductivity. The latter terminology refers to possible antenna applications of the quantum effects in superconductors. There is an exciting possibility that unusual and innovative antenna-like devices can be developed through use of such effects. These devices could be extremely small and yet very effective. They would not be subject to the electromagnetic theory of conventional antennas and would not suffer from the limitations imposed on the latter.

The previous paragraphs summarize, then, the motivations for the study which is reported here. We will begin by expanding on these ideas in a more or less descriptive way and by discussing associated topics. More detailed and quantitative analyses follow in later chapters. A logical starting point would be an itemization of objectives. The ultimate objective was to complete a study of electrically-small superconducting antennas. Some specific topics which may be viewed as sub-objectives of the major objective are as follows:

a. Miniaturization of superconducting antennas.
b. Effect of physical shape upon the Q of a superconducting antenna.
c. Long range magnetic coupling to a superconducting antenna.
d. Maximum radio frequency signal level of superconducting antennas.
e. Methods of coupling to superconducting antennas.

f. Materials and structural technology for superconducting antennas.

g. Antenna applications of quantum effects in superconductors.

As the work progressed, it was found that some of the sub-objectives loomed larger in importance than others. Item (a) seems of maximum importance because it has become increasingly clear that, aside from what may develop in connection with the quantum properties of superconductors, the advantages of the superconducting antenna stem entirely (directly or indirectly) from the possibility of miniaturization. If miniaturization is not of extreme value in the system then a conventional antenna should be employed. In fact, one must be prepared to trade-off antenna sensitivity even though an electrically-small, superconducting antenna is used to obtain miniaturization.

Item (e) ranks second in importance because coupling losses cannot be tolerated since the performance of the electrically-small, superconducting antenna (again aside from "quantum effect antennas") can only approach that of the conventional half-wave antenna in the limit. It is very important to evaluate all the promising coupling methods in their practical form, not the idealized forms which may be physically unrealizable.

Item (c) treats the analysis of mutual inductance between two coils in a way that recognizes the physical spacing between the coils and the finite speed of propagation of an electromagnetic field. The theory predicts several interesting possibilities which could be of great value in connection with electrically-small antennas. The objective here is to define these theoretical concepts, show how they could apply to electrically-small antennas, and discover the circumstances under which practical implementation is possible. Superconductivity enters because it may permit a close approach to the ideal theory.

Item (d) relates primarily to the transmitting antenna and the various properties of known superconductors. Present superconductors that are useful for time varying currents do not remain superconductive if the current levels are excessive. The objective here is to investigate the limits. Again, it is possible to treat the problem on a theoretical as well as on a practical level. Practically, the prospects for high power transmitting antennas are still in question; however, new experimental evidence on Type

II high current superconductors has reduced an earlier pessimism.

Item (f) can be categorized as support information except for certain postulated but as yet undiscovered materials which would lead to highly innovative antennas. The objective in this case is to locate a body of information on all materials that would be useful in the construction of electrically-small, superconducting antennas and associated equipment. Mechanical, thermal, and electrical properties are especially important. Some construction details may need to be discovered. Details concerning the cryostat or Dewar find their place under this item.

Early in the study it became apparent that the examination of electrically-small, superconducting antennas would be incomplete unless consideration were given the the quantum properties of superconductors. The classical properties would perhaps allow a closer approach to what is predicted by the ideal antenna theory (which ignores the ohmic resistance), but the quantum properties could give rise to new, different, and unusual antennas. This opinion has been partially justified in the interim through experimental work done elsewhere by other investigators. An additional objective, then, has been that of identifying those quantum effects in superconductors which might be relevant to electrically-small antennas. Methods of application, qualitative and quantitative prediction of performance also are objectives under this item.

Item (g), then, is the high risk-high reward topic because if really innovative, electrically-small superconducting antennas or antenna-like devices are to be discovered, it will be in this area.

Item (b), as it develops is more properly a second-phase type topic. It is a sequel to experimental work on miniaturization. The item has received relatively little effort because the problem is found to be quite intractable when approached purely from a theoretical point of view. Since an experimental approach, which would involve construction and testing of numerous typical models, is somewhat expensive; and since the justification of such testing is contingent on the non-marginal advantages of miniaturization, it was felt important to complete an evaluation the latter item first. Accordingly, the objective relative to Q versus physical shape has been to achieve a qualitative or semi-quantitative understanding of this factor.

A few words should be included in these introductory remarks concerning the interpretation of antenna function in the reported study. In order to avoid the restrictions of too narrow an interpretation, we will be satisfied with the somewhat oversimplified view that defines the antenna as a transducer which links the circuit with free space.

We will recognize conventional antennas as being bilateral transducers, thanks to the reciprocity theorem, but that one usually thinks in terms of transmitting antennas and receiving antennas because these often differ in important ways when studied in the light of engineering applications. (There are, of course, many instances where the same antenna plays both roles). Parasitic antennas could be viewed as comprising a third category. One can present a plausible argument for considering the parasitic antenna as being simultaneously a receiving antenna and a transmitting antenna.

Non-conventional antennas may well turn out to be unilateral transducers. Nevertheless they deserve to be called antennas if we accept the transducer definition. Some qualification might be required relative to what is being "transduced". Our interpretation is broad enough to include information as an entity that can be transferred from free space to the circuit, or vice versa.

Antenna efficiency will be nearly the sole concern in the case of conventional, electrically-small transmitting antenna. In the case of the conventional, electrically-small receiving antenna, it will be shown that antenna noise is of primary importance when superconducting and non-superconducting, electrically-small antennas are compared.

The frequency range of interest will extend from the lowest possible frequency to approximately the UHF range. There are two reasons for not extending the investigation beyond this upper limit: First, microwave antennas already are small and efficient; little would be gained by additional miniaturization because at that point the refrigeration equipment, not the antenna, determines the limiting volume. Second, unavoidable losses caused by high frequency effects in superconductors reduce the advantages of us the latter.

A good deal of prior work has been done by other investigators in

connection with the elimination of ohmic resistance in tuned circuits by using superconducting metals. This prior work and certain concurrent work have given considerable encouragement to our study. One of the sections in this chapter will be devoted to descriptions of important examples of such prior and concurrent work.

1.2 The Electrically-Small Antenna

An electrically-small antenna has a physical extension which is small in comparison to the maximum free space wavelength transmitted or received by the antenna. Thus, an electrically-small antenna may be physically-large if the operating frequency is low. (Conversely, an electrically-large antenna may be physically-small if extremely high frequencies are employed.)

Electrically-small antennas are in common use. For example, nearly every pocket transistor radio uses an antenna, a magnetic dipole, which is in physical size only 10^{-4} wavelength. Short electric dipoles are often used for portable receivers in the VHF range. At very low frequencies, in the VLF and ELF ranges, it is practically a physical necessity to use an electrically small antenna. For example, at 20, 000 Hz the half wavelength is approximately 5 miles; one could, but ordinarily does not, use an antenna this long for a receiver. To employ an extreme example, suppose an antenna is required for the study of electromagnetic radiation in the region of 1 Hz, the half wavelength for this frequency is approximately 100, 000 miles; one cannot avoid the use of an electrically-small antenna here.

The conventional, electrically-small transmitting antenna is never satisfactory from the standpoint of efficiency because much of the generated power is wasted in heating the ohmic resistance. On the other hand, conventional, electrically-small receiving antennas often are very satisfactory. Sensitivity rather than power efficiency is the important factor in reception, and sensitivity is limited by noise. (Perhaps it would be more correct to refer to usable sensitivity because the electromagnetic energy incident upon the antenna, any antenna, may be a mixture of noise and the desired signal. This noise may be much greater than the thermal noise generated by the antenna ohmic resistance, it may be much greater than the noise generated by the various noise mechanisms in the passive and active elements of the receiver).

Accordingly, the receiving function of electrically-small, superconducting antennas requires a more painstaking analysis and the relative merits of such antennas are not quite so distinct. One thing has become very clear. It is not possible to make sweeping generalizations about electrically-small, superconducting antennas. If the question is asked, "Are electrically-small, superconducting antennas better than conventional antennas?" the answer would need to carry many qualifications with respect to the many facets of the problem. The question could not be answered at all until the questioner could define the type of conventional antenna, its anticipated electrical function, its desired operational characteristics and specifications, all possible trade-off factors, and just what he meant by the word "better".

For the radio frequencies presently in use, there is almost always some advantage to the physically-small antenna. When other operational factors require also that the radio frequency be low, the degree to which a physically-small antenna would become electrically-small often forces the design toward limitations involving antenna efficiency, antenna sensitivity, antenna bandwidth, and antenna directivity. It is not clear whether or not these limitations are fundamental or whether they are technological.

Theoretically, the electrically-small dipole, either kind, possesses an aperture or cross section which is independent of physical size. The antenna may become vanishingly small physically, yet its theoretical aperture remains a constant $0.12\lambda^2$ square meters (if wavelength λ is in meters). This interesting condition exists because the radiation resistance of the antenna reduces with antenna size as rapidly as the square of the induced voltage, so the power available from the antenna remains constant.

Practically, the ohmic resistance does not fall off as fast as the radiation resistance when the antenna size is reduced; consequently, for an electrically-small antenna the radiation resistance may be orders of magnitude below the ohmic resistance. The ohmic resistance dissipates nearly all the power and the efficiency, gain, and effective aperture decrease rapidly to a useless value. Moreover, there exists the problem of coupling to a vanishingly small resistance; this has given rise to most pessimistic views in the literature.

A bandwidth limitation exists in applications where the antenna must be tuned. A very short antenna is highly reactive. If the extremely low radiation resistance

is to be effective in the transmitting mode, it must be associated with a relatively large current; accordingly, the antenna reactance must be tuned out by a conjugate reactance if conventional power sources are to be used. The result is an unusually high-Q circuit and the bandwidth, defined by the ratio of band center frequency to Q-factor, is quite small. Naturally, a small bandwidth means a correspondingly low information rate.

An incremental dipole is not highly directive, although its directivity is only slightly less that of a half wave dipole. It is logical, therefore, that an electrically-small antenna will not be highly directive unless it becomes possible to build a super-directive array. This presents a technological limitation rather than a fundamental limitation because superdirectivity is possible in theory. (Some authorities would insist that superdirectivity is so impractical that the limitation is fundamental.)

Of course, lack of directivity is not a deficiency in all cases. There are numerous applications where omnidirectional antennas are required; in such instances the incremental dipole is on a par with other omnidirectional antennas.

We will proceed to show how, where, and to what extent superconductivity can improve the class of electrically-small antennas, even those which are only theoretically possible.

1.3 Natural Antennas

Nature abounds with examples of what might be called electrically-small antennas. Everyone is familiar with the absorption and emission of electromagnetic radiation by atoms. Even at optical frequencies one is dealing with an electrically-small radiator; for example, the sodium atom with a dimension of approximately 10^{-10} meter is a very effective radiator or absorber of electromagnetic radiation having a wavelength of 0.589×10^{-6} meter. The radiating element is thus only some 2×10^{-4} wavelength long. Experiments bear out that the aperture or cross section of the tiny antenna is in good agreement with the predicted value of $0.12\ \lambda^2$.

Now this is not particularly impressive because we are accustomed to these processes. It is much more interesting to reflect upon the fact that these same atoms, molecules, and nuclei are able, as individual elements, to absorb and radiate

electromagnetic energy in the radio frequency range. Take for example a single proton with dimensions of the order of 10^{-15} meter, lying in a 1000 Gauss magnetic field. This object, via the process of nuclear magnetic resonance, is able to absorb or to radiate photons of frequency 4.25 MHz. Here the element is only 10^{-17} wavelength in size and yet it has an effective aperture that enables it to intercept this low frequency radiation.

The famous 1420 MHz radiation (or absorption) by neutral hydrogen exemplifies once more that the efficient electrically-small radiator is realized in nature. In this latter example the d/λ ratio (ratio of antenna length to radiated wavelength) is of the order of 10^{-10}. These electrically-small "antennas" are used to advantage in the atomic hydrogen maser [1] and they form the transmitting antennas for the radiation observed by some radio telescopes.

How can these tiny entities present such huge cross sections to the right frequencies? Somehow nature is able to realize the size independent aperture of the incremental dipole in molecular, atomic or nuclear resonances. When the radiative transition is viewed as a resonance, one is able to appreciate the behavior a little better, especially after observing the large cross section of an exceedingly high-Q tuned circuit which is exposed to the environment. In this sense the atom is analogous to an LC circuit tuned to the particular frequency and the long range coupling point of view may carry plausibility here.

We do not wish to imply that the behavior described in the various examples given above is of great mystery. Much theoretical work has been done in the field of quantum mechanics where the problem is handled in terms of transition probabilities. Even there, however, unanswered questions and differences of opinion exist. Grimes recently has given an interpretation to the problem that connects the resonant large scale radiator and the atomic radiator by more than just analogy [2, 3]. If this is so, we may be able to learn sonething from a study of these natural antennas that would be of benefit to us. Thus far the major clue lies in the high Q-factor that may be associated with such atomic-scale transitions; similar high Q-factors may be achieved via the use of superconductivity.

There is within our natural experience one extremely low frequency, incremental magnetic dipole. We speak of our own earth. It is widely supposed that the magnetic field of our planet is caused by large currents circulating in the liquid nickel-iron core. A most popular theory attributes these persistent currents to a magnetohydrodynamic mechanism where the earth is viewed as a self excited dynamo [4]. This is a very good theory provided some explanation can be found for the original exciting current which generated sufficient magnetic field for the build-up process. Why, also, does the field stabilize? (A self-excited dynamo stabilizes because of the saturation characteristics of its ferromagnetic circuit, otherwise it would destroy itself, given sufficient mechanical input.) Finally, how could an alternating field be produced? Some authorities believe, on the basis of recent evidence, that the field alternates every half million years or so. [5]

At any rate there is a long period fluctuation of the earth's field caused by internal current variations; that makes the earth a radiating, electrically-small dipole.

1.4 Superconductivity

This is not the place to present any comprehensive survey of superconductivity; besides, it is assumed that the reader has at least some knowledge of the subject; if not, there are many excellent, well written, and readily available sources for such information [6, 7, 8]. Some cursory discussion is appropriate, however, in order to highlight those salient features of the phenomenon which are relevant to this study.

There is a class of materials (consisting mostly of certain metallic elements and alloys) which has the following property: When the temperature of a particular material is reduced below a value characteristic of that material, its ohmic resistance decreases in a more or less discontinuous fashion to a value that cannot be distinguished from zero by any known technique of measurement. These "critical temperatures" range from a small fraction of a degree away from absolute zero to slightly above 18 deg-K, depending upon the material. It is this behavior which caused the discoverer of the phenomenon (Kamerlingh Onnes, 1913) to give it

the name "superconductivity".

The elimination of ohmic resistance means the elimination of Joule losses. This is the feature which suggests attractive possibilities for increasing the efficiency of transmitting antennas and for enhancing the relative magnitude of the radiation resistance in receiving antennas.

Unfortunately the property is marred by an undesirable trait: When the material is placed in a sufficiently intense magnetic field, superconductivity is destroyed. This includes the self-generated magnetic fields caused by currents flowing in the superconductors; consequently, there is a limit to the magnitude of the current that can be carried by a superconducting wire.

The maximum radio frequency current level is determined by the above behavior and some limitation will exist for the power handling capability of superconducting transmitting antennas. Superconducting receiving antennas will not be adversely affected.

A relatively new class of superconductors, known as Type II or hard superconductors, exhibits superior high current and high field characteristics. In the past it was thought that these did not have a sufficiently good a.c. response to be useful for service other than the d.c. electromagnets where they have demonstrated superb practical performance. Very recent experimental evidence confirms the possibility for radio frequency applications.

There is an a.c. effect in all superconductors where, at sufficiently high frequencies, the superconducting behavior disappears. Fortunately, such frequencies for the Type I superconductors are far above any contemplated in the proposed applications. The slight a.c. effects at the latter frequencies are completely negligible in comparison with other effects that may produce limitations.

Beyond the properties listed above lie a number of interesting effects which could lead to antenna applications that will be called non-conventional in the subsequent sections. Some of these effects are related to the magnetic behavior of superconductors. Solid Type I superconductors expel magnetic fields and multiple connected shapes (like rings) trap any magnetic flux that was contained in an opening while the material was still normal. Further, the trapped flux cannot take on any

arbitrary value but exists only in quantized levels [9]. This latter behavior is but one of a number of quantum effects. The popular literature gives interesting and readable accounts of the various effects which find partial theoretical explanation in the modern microscopic theory of superconductivity [10, 11].

It is fairly safe to say that superconductivity is not yet a closed subject [12] and that the discovery of new effects may be expected at any time.

Further discussion of specific effects will be included as it is needed throughout the report.

1.5 High Frequency Behavior of Superconductivity

One property of superconductors must be singled out and given special attention. While the resistance vanishes for steady currents, there is a finite resistivity to alternating currents. It is quite understandable that this would be a matter of great concern in view of the proposed application. Fortunately, the a.c. resistance still is vanishingly small for many of the superconducting materials considered for receiving antenna applications where the frequencies are below the microwave values. When Type I, or soft superconductors are employed, the a.c. resistance is so small that its effects are orders of magnitude below other effects that set limits to the usefulness of superconductivity in electrically-small antenna development.

The story is somewhat different for the case of Type II, or hard superconductors. The new niobium-zirconium alloys are typical of such. These materials, which resist the effects of magnetic fields developed by their own currents, are potential high current superconductors-as has so successfully been demonstrated in the case of high field superconducting solenoids- - and would seem logical choices for transmitting antennas; however, some investigators found that apparently such superconductors were severely frequency limited [13] and theories were developed to support this observation.

It cannot be overemphasized, and especially in this study, that at the present point in time superconductivity is not a completely understood phenomenon. It is unwise to speak with finality on any phase of the topic. Very recent developments

in connection with the Type II superconductors reinforce the latter assertion. Concurrent work by other investigators is even now beginning to indicate that the frequency limitation for alloy superconductors may not be so severe as originally was believed. The most encouraging work in this area has been reported by Cummings [14] whose experiments indicate that the frequency response of such alloys may extend to 600 MHz and higher.

1.6 Practical Realization of High Frequency, Superconducting Circuits

There is a considerable body of work reported in the literature related to the application of superconductivity to engineering electromagnetics. Harvey's Microwave Engineering [15] and Newhouse's, Applied Superconductivity [6] give sufficient references to start one on a path leading into all corners of the collection. It is a rapidly growing area of research and the reports of significant work are being published at a rate which makes it difficult to assess with any assurance the current state of the subject.

There are certain instances of prior and concurrent work that are felt to be of particular significance in this study; we discuss them briefly in the following paragraphs.

Nikola Tesla, in the year 1900, was aware that the Q of an LC circuit could be improved by reducing the ohmic resistance of the inductor through refrigeration; liquid air was mentioned in a patent issued to him in 1901 [16]. He spoke of the value of this idea in systems of transmitting intelligence to distant points. Actually, he refrigerated not what would amount to the antenna of the system, but rather the matching network of the antenna. He seemed unaware of the concept of radiation resistance and its part in radiation theory. His thoughts came before the discovery of superconductivity by Kamerlingh Onnes in 1911 so that his work applies only to the effects of reducing ohmic resistance and does not bear the larger implications inherent in the superconductivity phenomenon.

Work done on actual superconducting resonators is best exemplified by the cryogenic radio frequency tuner developed by Arams, et al. [17], under

Signal Corps Contract DA36-039-AMC-03321 (E). A lumped LC resonant circuit was constructed which, in the superconducting mode, exhibited Q's ranging from 350,000 to 600,000 over the 6.3 to 21 MHz tuning range. Other experimental circuits yielded unloaded Q's as high as 2.2 million near 20 MHz at 4.2°K. These circuits were used only as narrow band RF tuners. They, necessarily, were shielded by a superconducting shield and were not exposed directly to radiation fields.

Their work is of great value to us because it provides some of the experimental data which, it was thought at the outset of our study, would be needed to lend assurance to various predictions. Some specific points of concern which have been settled by their experimental work are as follows:

1. It is possible, at frequencies in the neighborhood of 30 MHz to eliminate the ohmic losses in a conductor and thus realize high-Q inductors. Unloaded Q-values of 2×10^6 have been demonstrated at these frequencies and the evidence seems to indicate that still higher Q-values can be provided at higher frequencies. These Q-values were obtained for coils contained in a superconducting shield.

2. The Q-value of 2×10^6 represents the practical realization of a theoretically predicted Q-value of 770×10^6. The reduction in value is believed due to dielectric losses in the insulating materials which support the structures. This data, along with the structural details presented in the referenced work, will permit realistic estimates of limitations to be expected in superconducting antennas.

3. It is shown, by quantitative data, that high stability can be expected in a properly constructed superconducting tuned circuit. The demonstrated stability, better than ±0.2 db for a 30 minute run, is sufficient justification for confidence in the stability of superconducting antennas.

4. The prospect for high-Q, high power, superconducting transmitting antennas using thin layers of Type I superconductors seems dim in

view of the experimental data. Power levels of the order of 0.1 watt apparently caused quenching due to some undetermined cause. The theoretical power limit for the mode of operation is three orders of magnitude higher. It is suspected that surface or geometrical properties of the superconducting material are involved in this limitation so that some improvement could be expected.

5. The extremely low insertion loss of the superconducting preselector demonstrates that the calculated effectiveness of magnetic loop coupling is a genuine physical possibility. Under some conditions of operation the entire insertion loss could be traced to losses in the non-superconducting input and output connectors.

An actual electrically-small, cryogenic antenna has been reported by More and Travers [18]. The 20 MHz radiation from a 12-inch monopole with loading coil was observed while these elements were cooled to 77° K and then 4.2° K. The radiation efficiency at 77° K was 2.5 times that measured at 290° K. The radiation efficiency increased to 4.3 times the 290° K value when the elements were cooled to 4.2° K.

The matching network was not cooled, the report (a short note) fails to give details on materials, and the authors' interpretation of superconductivity is not clear; nevertheless, these results are in definite support of the feasibility of achieving an efficient electrically-small antenna by utilizing superconductivity.

The work by Nahman, Gooch, and Allen on superconductive coaxial transmission lines and superconductive delay lines [19], [20], [21], is relevant to any study of superconducting antennas. It shows that a spectacular result can be achieved in radio frequency circuits by using superconductivity to eliminate or greatly reduce ohmic losses. It gives valuable theoretical and experimental details and data on radio frequency properties of insulators at cryogenic temperatures. Separation of losses was possible in the experiments and the dielectric losses are well defined. The high frequency (microwave range) limitations of superconductors are clearly evident in the data.

One encouraging experimental result has a bearing on the maximum R. F. signal level that can be handled by a superconductor. It was found that the transmission line could handle a current as high as half that predicted by theory. This is in sharp contrast to the observations described above in connection with the R. F. tuner. A thin superconducting layer was used in the case of the tuner whereas a solid superconductor was used for the center conductor of the line.

Their research has been reinforced and extended by Cummings and Wilson [22] who conducted high power pulses through superconducting transmission lines. The typical transmission line consisted of subminiature coaxial cable with the following mechanical characteristics:

Inner Conductor	0.015-in. o.d.
Primary dielectric	0.026-in. wall FEP TEFLON *
Outer Conductor	Braided 0.007-in wire

Through approximately 300-feet of this they transmitted 15 kilovolt, 300 Ampere pulses of 2.5 microsecond duration and 2 nanosecond risetime.

The inner and outer conductors were fabricated from a Type II superconducting alloy, specifically, Nb-25% Zr.

There are two factors which make this experiment so significant. First, these high power pulses were transmitted without quenching superconductivity. Second, the Type II superconductor, which commonly is thought to be severely frequency limited, handled frequency components (as implied by the stated risetime) as high as 180 MHz.

The secret seems to be an inherent delay in the quenching of superconductivity until a certain amount of <u>energy</u> (the value being related to the energy gap characteristic of the superconductor) has been accumulated. The superconductor integrates the power until that energy value has been reached before it "goes normal". In other words, a high power pulse can be transmitted in the superconducting mode provided the pulse duration is sufficiently short.

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If we extrapolate from these interesting experimental results, it would seem that short, high power, high frequency bursts could be handled by an electrically-small, superconducting, transmitting antenna fabricated from these materials.

1.7 Conventional Application of Superconductivity To Electrically-Small Antennas.

We wish to establish a distinction between the various ways of applying superconductivity to the development of antennas and antenna theory. It has been pointed out that superconductivity embraces much more than its name implies. The vanishing resistivity of the conductor is only a small part of the phenomenon, as we now know it. The magnetic behavior and the quantum effects are additional features that must be considered. Other effects and properties are certain to come to light in the future because the theory of superconductivity is far from being complete.

Let us define the conventional application of superconductivity to electrically-small antennas as that part of the study which deals with the advantages to be gained by eliminating the ohmic resistance of conductors used in the antenna structure. Having identified these advantages we must also determine the degree to which they can be realized in the light of various effects which impose limitations on the range of that property.

Under non-conventional application we will treat possibilities created by the more subtle aspects of superconductivity.

The conventional application of superconductivity to electrically-small transmitting antennas results at first glance in a more efficient antenna because no power is wasted in heating the ohmic resistance of the conductors. A much more efficient antenna is expected because the ohmic resistance may be considerably greater than the radiation resistance. However, we must be quick to limit and qualify this seemingly obvious conclusion. The impedance of the antenna must be taken into account, it is frequency dependent; and, the energy source must be considered. There may be also some reason to allow for what is being transmitted: intelligence,

energy, a reference frequency? The concept of efficiency involves the entire transmitting system and its purpose.

It is easy to pinpoint the crux of the larger problem; it is the series reactive element of the antenna. Any current which flows in the radiation resistance must flow through the series reactance, there is no way out of it. (We are using the circuit equivalent for a radiating element). In the case of an electrically-small antenna, it is well known that the series reactance (capacitive for the electric dipole, inductive for the magnetic dipole) is much larger than the radiation resistance; in fact, it is much larger than the sum of the radiation resistance and the ohmic resistance in most presently useful electrically-small antennas. Consequently, a large voltage must be applied to the antenna to obtain even a moderate current in the radiation resistance.

If the ohmic resistance is negligible, the large reactive voltage does not in itself signify a poor efficiency because a reactance does not consume average energy; however, it may impose technologically difficult or impossible conditions on the energy source. For example, it would be easy, at this moment, to design a 99-percent efficient, 1-Watt, electrically-small superconducting transmitting antenna provided someone could supply a 50-kilovolt transmitter with a series impedance less than 100-Ohms such that the resistive component of this impedance were less than 10^{-2} Ohm. (These latter conditions would not need to hold beyond 1-Watt). Moreover, if the transmitter could approximate such performance over a wide frequency band, the antenna would give correspondingly efficient performance over that same frequency band. Thus, since the burden of the problem can be shifted easily to the transmitter, we see that the study of electrically-small, superconducting, transmitting antennas must involve the entire transmitting system.

Such transmitters are not available (however, see section 7.1) we must do the best we can with conventional equipment; accordingly, the first thought is to resonate the antenna reactance so that a low voltage, low impedance transmitter can be used. But now we have a severe bandwidth limitation and, while the system may be efficient in the utilization of energy, it may not possess a sufficiently high information rate. The latter limitation may, however, not be of concern if the purpose

of the system is to transmit energy at a single frequency.

This then is a cursory survey illustrating the conventional application of superconductivity to electrically-small transmitting antennas and the basic problem that will be encountered.

Efficiency is of relatively little concern in the case of receiving antennas; not altogether so, because a receiving antenna may be part of an energy transmission system. Normally however, it is information that is to be transmitted and received; at the receiving end the power level probably is so low that efficiency becomes completely irrelevant to the problem. In information transmission, the signal to noise ratio is the all-important property when the received signal is feeble.

Viewed in this light the analysis seems to lead to the following conclusion: The conventional application of superconductivity to the electrically-small receiving antenna cannot result in performance that is superior to the performance of a non-superconducting half-wave dipole.

The basic logic is this: The ultimate figure of merit for the conventional antenna is its signal to noise ratio. But the ideal electrically-small antenna has the same available signal power as the non-superconducting half-wave dipole because the apertures of both are practically the same. Moreover the available noise power, KTB, is the same for both in the same system bandwidth. Therefore, the ratio of available signal power to available noise power, is the same for both.

In interpreting this "logic", it is important to understand that temperature T in kTB is <u>not</u> the ambient temperature of the antenna structure for ideal antennas, it is the temperature of the radiation resistance. These two temperatures are quite independent, a conventional antenna may have an ambient temperature of 290 deg-K while, at the same time, the effective temperature of its radiation resistance may be 10,000 deg-K. And, suppose a superconducting antenna has an ambient temperature of 4 deg-K. That has no effect on the temperature of its radiation resistance which could be 10,000 deg-K.

Within the restriction described above there still remain instances where the elimination of ohmic resistance by superconductivity is advantageous. Probably the only real advantage of the conventional application of superconductivity

to receiving antennas is the possibility of efficient miniaturization. There are direct and indirect benefits in such miniaturization. The direct benefit is rather obvious; the indirect benefits stem from the fact that the efficient miniature antenna approaches a truly concentrated or point source of dipole radiation. This latter is important when the elements are to be used in an array or near parasitic elements that are deliberately placed or which may be unavoidable. With a point source the interaction of elements is electromagnetic and is not perturbed by the electrostatic coupling which takes place between the large conventional elements.

It is felt that this latter property, the point source behavior, may reopen some thinking in regard to superdirective arrays.

Progressing with the realization of these basic facts, it now becomes necessary to discover and explore all the direct and indirect benefits of miniaturization. Also, it becomes necessary to concern oneself with the interface between the antenna and the rest of the system. We cannot afford to squander any of the signal to noise ratio in a lossy coupling network or an unsuitable detector.

1.8 Non-Conventional Applications of Superconductivity To Antenna-Like Functions

All other possibilities for improving electrically-small antennas by using the properties of superconductivity are relegated to the category of non-conventional applications. The previous section demonstrated that plausible arguments could be given in support of fairly definite limits on what can be accomplished by eliminating ohmic resistance in electrically-small transmitting and receiving antennas. These limits map out a closed area of study and there is reasonable assurance that all the interesting possibilities can be identified.

These limits do not apply to what may be possible when the other properties of superconductivity are brought into play. One can still anticipate the appearance of really innovative electrically-small antennas, transmitting as well as receiving antennas. It will be necessary to remain open-minded about the definition of "antenna" because some of the properties may permit only a unilateral function;

however, if one accepts the definition, "An antenna is a transducer between the electric circuit and free space, allowing energy and/or information to be transferred from one to the other," the unilateral behavior is included.

Primarily it is the quantum effects in superconductors that we have in mind we speak of non-conventional applications but this second area is not exclusive of other interesting effects which are not directly associated with the infinite conductivity property.

1.9 Natural Cooling

It has been said that the advantages of superconducting antennas would be particularly meaningful in the space environment because the natural low temperatures would obviate the need for refrigeration equipment. Let us examine the possibility of achieving superconducting temperatures by this means. If natural cooling is feasible then electrically-small superconducting antennas may have some place in space exploration. Any advantages would be enhanced greatly and the frequency range of applicability would be extended to microwave values because the minimal volume of the antenna system would not be determined by the volume of the cryostat and associated equipment, as it is in high temperature environments.

The temperature requirements are such that 18 deg-K must be reached, and it is preferable that values as low as 5 deg-K be attainable.

In space, heat is transferred by radiation; accordingly, we must study the radiation balance between the antenna and all the significant radiators in the vicinity. It turns out that the minimum obtainable temperature may be 3 deg-K because the universe itself is suspected of being a 3 deg-K black body.

In the solar system, the heat input to an isolated antenna is determined by the solar constant at its location. At the sun-earth distance the solar constant I_e is approximately 0.14 Watts per square centimeter. The amount of power absorbed depends upon the absorptivity of the object and its projected cross section. Absorptivity α varies from zero for perfect reflection to a maximum of unity for a black body. While it is possible to approach unity with practical surfaces, the practical

minimum obtainable value (for solar radiation) is approximately 0.1.

For a simple surface the heat input would be given by

$$H_i = \alpha I A_n \tag{1-1}$$

where I is the Watts per square centimeter incident radiation density and A_n is the projected surface having absorptivity α. If I is due to solar radiation, the object being distance r from the sun, then its value can be computed via

$$I = 0.14 \, (r_e / r)^2 \text{ Watts per sq. cm.} \tag{1-2}$$

where r_e is the sun-earth distance. A more complicated projected surface having various α's would be handled by an appropriate summation or perhaps by an integral.

If the object is a fairly good conductor of heat, as antenna most likely would be, the radiated heat is given by the integral

$$H_o = \oiint \alpha \sigma \, T^4 dA \text{ Watts} \tag{1-3}$$

where the Stefan-Boltzmann constant $\sigma = 5.67 \times 10^{-12}$ Watts per square centimeter per deg-K^4, T is the absolute temperature in deg-K, and the integral is over the entire surface of the object.

For a radiation balance $H_i = H_o$ and to achieve low antenna temperatures the strategy would be to present a small, poorly absorbing cross-section to the radiating source and a large surface with a high absorptivity to empty space.

Let us now apply these facts to simple but meaningful examples. First, the ratio of the radiating to absorbing surface will be high in a successful application. Let the ratio be designated by K. The minimum practical absorptivity of 0.1 will be used for the absorbing surface and the remainder of the structure will have a black body surface. The antenna structure will have a uniform temperature T. Under these conditions, equating expressions (1-1) and (1-3) yields

$$T = \left[(1.76 \times 10^{10}\, I) / K \right]^{1/4} \text{deg-K} \qquad (1\text{-}4)$$

where we are neglecting the relatively trivial amount of radiation from the small fraction of the surface which absorbs the incoming radiation. Assuming that the antenna is at the sun-earth radius (but not anywhere near the earth) and that K = 100, we obtain a natural temperature of 71 deg-K. If the antenna is moved out to $10r_e$ the temperature will fall only to 22 deg-K; it can be seen that direct exposure to the sun cannot be tolerated.

Let us try a radiation shield; place the antenna at the sun-earth radius but very remote from the earth. Position a highly reflecting sphere, say $\alpha = 0.1$, between the antenna and the sun. It will probably be impossible to maintain a K-factor of 100 if the sphere is at a practical distance; let the K-factor drop to 10 with the antenna located 5 sphere diameters behind the shield. Under these conditions the radiation balance expression and the inverse square radiation law show that the antenna temperature will become 16 deg-K.

These simple calculations show that natural cooling is indeed feasible even with crude shielding techniques. Much more can be done with sophisticated shields; for example, cascaded shields would be more effective than single units. It is of course not necessary to construct the radiation shields from conducting materials.

1.10 Effects of Antenna Environment

In almost every case it is impossible to really identify the total antenna of an operating system. The local environment of what we normally point out as being the antenna often plays a great part in the function of that device. For example, a television antenna may be defined through performance specifications obtained in a free space chamber, but when the antenna is placed into actual service, the presence of guy wires, nearby structures, electrical wiring under the roof, power and communications transmission lines, other nearby antennas, and even transient objects such as vehicles and aircraft may make a shambles of those specifications as they relate to this particular antenna. We cannot ignore the problem; it is there and

the total antenna is somewhat ambiguous.

The problem is not quite so severe in the case of highly directive antennas; however, our antennas will be fairly omnidirectional because they are electrically-small. It will be difficult if not impossible to avoid having objects in the near field; in many cases such objects will behave like parasitic elements, thus altering the radiation pattern, and in doing so will become part of the antenna system. Consequently, we may have a superconducting driven element but we do not necessarily have a superconducting antenna system because these parasitic elements are not superconductors.

Such behavior affects both the transmitting antenna and the receiving antenna. All the energy dissipated in the parasitic elements must come from the transmitter, it is the source of the induced currents and these currents are apt to be relatively significant in nearby objects to which the inductive and/or capacitive coupling is strong. If the antenna terminals are connected to an impedance bridge, that portion of the measured series resistance which is greater than the radiation resistance of the superconducting element is due primarily to the parasitic elements and other energy absorbing material in the near field of the driven element.

These same extraneous resistance elements are responsible for the ultimate limitation of the receiving antenna. They are the noise producing resistances in the antenna circuit. If any of the sources are contained in the cryostat, their noise contributions will be minimal because of the exceedingly low temperature; however, most of the parasitic elements and other media will be external to the low temperature bath, they will be represented by hot, noisy resistances.

A quantitative assessment of the effects of antenna environment is vital to any study of electrically-small, superconducting antennas. Sections 6.5, 6.6, and 6.7 of this report deal with the effects of antenna environment.

1.11 Innovations In Superconductivity

In this section we attempt to give a brief survey of those past, recent, and current investigations in superconductivity which are somewhat innovative in nature and which are relevant to the application of superconductivity, in the broadest sense of the word, to electrically-small antennas.

One most elusive goal is the development of the high temperature superconductor. In our case the need for a cryostat or Dewar detracts greatly from the possibility of miniaturization. A minimal volume is required for a practical cryostat; consequently, nothing or little can be gained by miniaturizing an antenna which, in its normal form, already is as small as the cryostat.

There is no real assurance that a room temperature superconductor can be achieved. Superconductivity is one of the least advanced areas of solid state theory and much progress will be required before the theory attains the degree of certainty present in something like semiconductor theory. At present the theorists are divided into two camps. One group sees the possibility of high temperature superconductivity in its interpretation of existing theory; the other group, equally eminent, rejects the idea [23].

Parmenter [24] sees in the theory a prediction of the superconducting state at high current densities, even in metals that are not now classified as superconductors, and even at temperatures as high as normal room temperatures. Little [25, 26] has discussed the possibility of synthesizing an organic superconductor which would operate at room temperature, there are indications that the proposal has been taken seriously by industry [27]. Okress [28] has suggested that the phenomenal sensing ability of certain insects may be due to sense organs that behave like superconducting (organic) antennas. At the time of writing, a Government-sponsored program is being conducted jointly at the National Bureau of Standards' Institute for Materials Research and the U. S. Navy's Marine Engineering Laboratory with the objective of producing alloys that become superconducting at higher temperatures than are now possible.

1.12 Summary

Can superconductivity be used in the development of better electrically-small antennas? This first section has been an introduction to the report of a study which attempted to answer that question. The initial survey began by considering natural electrically-small antennas ranging from an atom to the earth itself. Some of these natural antennas are excellent electrically-small radiators and absorbers of electromagnetic energy; they serve as models and in a sense they represent a challenge. In many instances, the mechanisms supporting the operation of the natural electrically-small antennas seem to bear some analogy to the phenomenon of superconductivity.

The principle argument for applying superconductivity to electrically-small antennas is that the ohmic resistance can be eliminated thereby. The ohmic resistance is not a really serious problem in electrically-large antennas because the radiation resistance is much greater. In the electrically-small antenna the radiation resistance is much smaller; consequently, the transmitting antenna becomes very inefficient and the receiving antenna becomes more noisy, in a relative way. The ohmic resistance <u>can</u> be eliminated and experimental evidence to that effect was cited.

Since there is no question about the possibility of eliminating the ohmic resistance of electrically-small antennas in all radio frequency ranges where such antennas are important, the next step involves a realistic assessment of the improvement which is possible. Is there a striking improvement or is the improvement marginal? The reader was left in doubt on that point at this time, at least no sweeping generalizations were made. Criteria were mentioned and there was some indication that each application must be considered separately.

It should be possible, at least, to achieve closely what is predicted by classical antenna theory for the electrically-small antenna; however, the usefulness of such antennas may be limited because of what is available in existing transmitters and because of natural environmental noise in the frequency ranges of interest.

Superconductivity is more than a phenomenon which eliminates ohmic resistance. There are other effects which may be useful in conjunction with

electrically-small antennas. These non-conventional applications may become of greater importance in the future and may lead to really innovative antennas. Section 9 will go into detail on this topic.

Even now, it can be seen that the application of superconductivity to electrically-small antennas is an open ended problem if all aspects of superconductivity are to be taken into account. The problem can be approached on a number of levels, ranging from one of purest abstraction to a level dealing with present technology and needs. There has been an attempt in the subsequent material to break the problem into at least three divisions, although these are not presented formally as such: It was felt important to determine what is theoretically possible, that is, what are the fundamental limitations if any? Secondly, what possibilities depend upon advances in other areas? Finally, what is possible now?

SECTION II

MAXIMUM RADIO FREQUENCY SIGNAL LEVELS

2.1 Transmitting Antennas And Receiving Antennas

Ordinarily the maximum radio frequency signal level of a superconducting antenna will be reached when the self-generated magnetic field attains the critical value, beyond which superconductivity is destroyed. The self generated magnetic field depends in turn upon the currents flowing in or on the antenna conductors. Finally, the critical magnetic field value varies widely with the superconducting material.

One would imagine that the application of superconductivity to receiving antennas would not be affected seriously by these considerations. Except for a few unusual applications, superconductivity would be used to enhance sensitivity in some direct or indirect way. This implies very small radio frequency signal levels and even if there is a large dynamic range in received signal level, the superconducting mode certainly would not be required at the high signal level.

The exceptions could occur in those cases where it would be necessary to maintain an extremely high Q-factor at moderate signal levels because very large currents can flow in high Q circuits excited by relatively small signals.

When extremely high Q-factors are employed, there can be limitations due to the high electric field intensities which are generated even though the currents are less than the critical values. In a simple LC tuned circuit the expression $V_C = V_L = QV_S$ relates the voltage across either of the reactive elements to the series exciting voltage. A relatively small series voltage can cause significant voltages in circuits where a Q-factor of 10^6 or higher is possible.

In the case of electrically-small, superconducting, transmitting antennas where the currents are arbitrarily large, the maximum power level is

limited either by the critical current or by dielectric breakdown. It is quite possible that dielectric breakdown could occur before the onset of the critical current limitation because an electrically-small antenna is by nature a high-Q device; however, this sort of failure is not a fundamental limitation. Large surfaces and sufficient insulation can prevent dielectric breakdown in most cases, up to the point where the antenna attains such a bulk that the advantage of miniaturization is lost. The most likely failure mode is one where the critical current is exceeded.

It is easy to see that large antenna currents are required for even trivial amounts of radiated power when one is dealing with an electrically-small antenna. The radiated power is given by the product $R_a I^2$ where R_a is the series radiation resistance of the antenna and current I is the series antenna current. For most of the antennas we have in mind, the radiation resistance can be much less than 1 Ohm; consequently, antenna currents of the order of tens of Amperes may result in radiated powers of the order of only one Watt.

The question in regard to maximum signal levels attainable in superconducting antennas seems then to be directed primarily toward the class of transmitting antennas. We do not need to be overly concerned with the problem in the case of receiving antennas unless we are dealing with exceedingly high Q-values.

Parasitic antennas are, in a sense, both transmitting and receiving antennas, simultaneously; however, these devices usually are part of a transmitting or a receiving system and the currents will be of a corresponding order of magnitude.

2.2 Silsbee's Rule

Superconductivity can be destroyed or "quenched" by the application of an external magnetic field. The minimum magnetic field required to produce this behavior depends on the geometrical shape and on the orientation of the superconductor in the field. The defined critical field H_c is the minimum magnetic field necessary to quench superconductivity in a long right circular cylinder of the particular material, the cylinder having its axis oriented parallel to the field.

For many superconductors the critical field varies with the absolute

temperature of the specimen in accordance with the empirical relation

$$H_c \approx H_0 \left[1 - (T/T_c)^2 \right] \qquad (2\text{-}1)$$

where T_c is the critical temperature of the material. The expression is meaningful only for temperatures less than the critical temperature. It is seen that the maximum critical field is tolerated at zero absolute and that no magnetic field can be tolerated if the specimen is near its critical, or transition, temperature.

Now, a current carrying conductor generates its own magnetic field. A long straight conductor has a field at radial distance r given by the well known expression (MKS units)

$$H = I/(2 \pi r) \qquad (2\text{-}2)$$

and the field direction is normal to the plane containing the wire and the radius vector corresponding to r. For a round wire the expression gives the field at the surface of the wire when r is the wire radius.

It was predicted by Silsbee in 1916 that superconductivity would be quenched in such a wire when the surface field due to the current exceeded H_c. This prediction was confirmed in the case of soft superconductors provided the wire was not too thin (0.1 millimeter still is thick), neither does it hold if the superconductor is in the form of a thin film. The critical current is less than the predicted value for both of the latter instances.

Order of magnitude calculations show that the critical currents in useful cases are in the neighborhood of 10 Amperes. For example, a 1-millimeter diameter wire carrying 25 Amperes will generate a surface field of approximately 100 Gauss. A Type I superconductor could carry currents of this magnitude if it were in the form of a straight wire; however, if a coil were used, it would be necessary to allow for the superposition of the magnetic fields.

Currents of this magnitude while seemingly appreciable are not really

so because of the low radiation resistance of the electrically-small antenna. The antenna in question might well have a radiation resistance of 10^{-4} Ohm, in which case the radiated power would be only 63 milliwatts. (The same antenna in the normal state could easily have an ohmic resistance of 0.1 Ohm so that 63 Watts of power would be dissipated in heat as the 63 milliwatts are radiated. The use of superconducting material eliminates the heating loss).

Antenna currents that are one or two orders of magnitude greater will be required for transmitting purposes; unfortunately, the critical fields of Type I superconductors are too low to permit such values, and the prospect is not good for transmitting antennas constructed from these materials. The possibility must not be discarded, however, because there are unanswered questions regarding the transient behavior of superconductors. Silsbee's rule pertains to d.c. operation and primarily to the soft superconductors. Current limitations under a.c. operation seem to be different, and the hard superconductors permit exceedingly high current densities which may be available to special modes of a.c. operation.

2.3 Existing Experimental Evidence

Arams, et al., [17] provide experimental evidence relating to radio frequency signal level limitations in elements using lead-plated surfaces. (Plating thickness is not available; however, it is stated to be considerably thicker than the 2-microinch superconducting penetration depth). Briefly, a high-Q (25,700) superconducting tuned circuit using lead plated inductor and capacitor absorbed only 88 milliwatts at a resonant frequency in the vicinity of 20 MHz before a quenching effect occurred.

At this signal level the peak current was calculated to be 1.3 Amperes, corresponding to a field of only 6 Gauss. For lead, one would expect to reach a coil current of 120 Amperes for the critical field of 550 Gauss at the liquid helium temperature of 4.2 deg-K. (Actually, this value may be optimistic in view of the superposition of field from nearby turns).

The investigators state that the exact cause of the quench is not

understood. It is stated further that an increase in power level to 18 Watts resulted in noise from the Dewar, probably due to voltage breakdown.

Moore and Travers [18] have radiated energy at frequencies near 20 MHz with a 12-inch whip antenna and loading coil both cooled to 4.2 deg-K. The radiation efficiency increased by a factor of 4.3 times the 290 deg-K value. The input power to the system may have been as great as 300 Watts.

Nahman and Gooch [20] conducted measurements on superconducting coaxial lines. The signal level was not particularly high, being some 17 volts into a 50 Ohm line. (The signal was in the form of a 1.7 nanosecond pulse with rise time of 0.4 nanosecond). However, the experiment was successful in that no measurable deterioration of rise time was experienced in a 100-ft length of the line and the small attenuation could be traced to input and output connections. The remarkable feature is the use of a 10-mil niobium solid center conductor. It is reported that a 3-mil conductor of the same material could carry a supercurrent up to 50-percent of that value predicted by the Silsbee hypothesis before going normal. (One assumes from the text that a d.c. measurement mode was employed in this determination).

One source of great encouragement is the experimental work of Cummings and Wilson [22] who, on the basis of their work, assert that high field superconducting alloys are capable of transporting high current densities at high frequencies. Some of the details were mentioned in section 1.6. The interpretation of their experimental results would lead one to believe that hundreds and even thousands of Amperes of current could be supported by a transmitting antenna made of a Type II superconductor if the signal were applied in the proper manner. Short pulses with nanosecond risetimes and enormous power levels can be supported in the superconducting mode by such materials provided the energy in the pulse does not exceed a characteristic switching energy. The switching energy is related to the energy gap of the superconductor, and power peaks in the megawatt range are indicated.

There is no information on pulse repition frequency and this leaves open the question of recovery time after the application of a pulse. If the recovery

time were short, and the proposed energy gap theory would seem to indicate quick recovery, it would be tempting to view the first half cycle of an r.f. burst as a pulse and the second half cycle with opposite polarity as a successive pulse, and so on. This failing, it would seem possible to at least think in terms of an r.f. burst with an integrated energy that is less than the switching energy.

2.4 Prediction of Limitations

At the beginning of this study there would have been an air of caution in the prediction of maximum radio frequency signal levels. The critical current limitations of Type I superconductors and the poor frequency response attributed to the high current Type II superconductors created some pessimism in regard to the usefulness of superconductivity in electrically-small transmitting antennas. A maximum power value of the order of one Watt (for a true, electrically-small, superconducting antenna) would have been appropriate for a predicted limitation. The best that could be hoped for by way of improvement was some practical implementation of the theory proposed by Parmenter [24] in connection with high current superconductivity.

The research by Cummings and Wilson has given a new perspective to the possibility of efficient, high power, electrically-small, superconducting transmitting antennas. More experimental research needs to be done in order to extend their work in an obvious direction. Short, high power r.f. bursts should be substituted for the pulses; recovery time should be investigated. The limitation may well be the described switching energy; although, the manner of interpreting switching energy in connection with an r.f. signal is not now clear. Until something is done by way of extending the cited research, there is no point in predicting limitations.

SECTION III

A FIGURE OF MERIT FOR RECEIVING ANTENNAS

3.1 Antenna Noise

We make a special point of discussing antenna noise because, shortly, it will be shown that noise is a primary factor to be considered in comparing receiving antennas. Two major types of noise must be taken into account: Thermal noise and sky noise. There are other incidental types of noise that are important when the antenna is in a special environment conducive to the particular type (for example, when the antenna is being bombarded by a flux of charged particles); however, we are not interested in these at the moment.

Thermal noise is the familiar Johnson noise generated by the random thermal motions of charge carriers in resistive circuit elements. In the case of a single isolated resistor R (Ohms) at absolute temperature T (deg-K), the root mean square noise voltage generated at its open circuited terminals in bandwidth B (Hz) is given to sufficiently good approximation by the expression

$$\overline{v_n} = \sqrt{4kTRB} \text{ Volts} \tag{3-1}$$

where k is Boltzmann's constant, 1.38×10^{-23} Joule/ deg-K.

There are a number of resistances associated with any antenna. It is important to identify all these resistive elements and to determine exactly how they appear in the antenna circuit. It is even more important to determine the value of the absolute temperature that should be assigned to each element because, normally, they are not at the same temperature. The three obvious resistances to be associated with a simple antenna are the ohmic resistance of the antenna conductor, the radiation

resistance, and the resistance of the dielectric material used to insulate or support the antenna structure.

The less obvious resistances enter the antenna circuit by capacitive or inductive coupling to objects or media in the nearby environment. Sometimes the coupling is direct, as in the case of a monopole antenna with a lossy ground plane. Sometimes matching elements or tuning elements form an integral part of an antenna structure and introduce resistive losses of various types. Sections 6.3 through 6.7 of this report discuss these various losses and effective resistances in some detail and point the way to quantitative determination of values.

It is very difficult to find a logical terminating point for the antenna because the antenna load plays such an important part in the total behavior. It is wrong to consider the antenna separately and then the receiver input separately because there is a great deal of interaction between the two. The dividing line between the antenna and the receiver is very indistinct; it is sharp only when the input impedance of the receiver is infinite, and even then the matching section, if any, should be considered part of the antenna circuit. Section 6.9 discusses antenna loading losses; sections 4.3 and 4.4 deal quantitatively with the antenna-receiver input impedance interaction. In our present discussion, since the receiver is arbitrary to a certain extent, it may be well to think in terms of an independent antenna assuming, therefore, an infinite receiver input impedance and in principle, a perfect receiver. Accordingly, at this point we ignore any thermal noise due to the receiver input impedance.

A calculation of thermal noise involves the determination of the total effective series resistance of the antenna, the effective noise temperature, and the noise bandwidth of the antenna. The problem of calculating the effective noise temperature of a group of resistors, all at different ambient temperatures, is discussed by van der Ziel [29].

Sky noise is more properly a signal that enters the antenna from the surrounding radiation field, thus if the antenna were placed in a perfectly shielding enclosure made from a material with infinite conductivity, the noise signal at the antenna terminals would be true thermal noise only. The developed sky noise in an

antenna is not reduced when superconducting materials are used, in fact, the available sky noise may actually be increased, again because sky noise is a signal.

It is common to include the sky noise contribution via the radiation resistance by attributing to the radiation resistance an effective temperature such that the thermal noise developed by a resistor at that actual temperature would be equivalent to the sky noise. If the antenna is "looking" into a black body radiation field, the radiation resistance takes on that black body temperature. The temperature may be very low in the case of a highly directive microwave antenna pointed into deep space. It will be in the neighborhood of some 300 deg-K for any antenna within a cave deep in the earth.

At the lower radio frequencies, beginning approximately at 1000 MHz, the received sky noise increases sharply due to contributions from what is called atmospheric noise and galactic noise. Hogg and Mumford [30] review this effect and show that down to about 20 MHz an expression equivalent to

$$T_a = \frac{2.6 \times 10^{19}}{f^2} \tag{3-2}$$

gives the average sky temperature to a good approximation. The temperature is in deg-K and frequency f is in Hertz. At 20 MHz this predicts an effective sky temperature of 65,000 deg-K. Measured values are in excess of 150,000 deg-K for antennas at that frequency pointed in the direction of the galactic center.

Maxwell and Stone [31] describe the natural noise fields from 1 Hz to 100 kHz. At 100 kHz the natural noise density is shown by measurement to be in the vicinity of 1 microvolt per meter per $Hz^{1/2}$. Available power density in the electromagnetic wave would be of the order of 10^{-15} Watt per square meter per Hz bandwidth. If this is converted into an effective temperature on the basis of a 1 square meter aperture, the result is a value of the order of 10^8 deg-K which is not too far from the approximately 10^9 deg-K value that would be predicted by the expression given earlier, were the latter valid at these low frequencies. Actually, the aperture of such a low frequency antenna would be much larger than a square meter in which case the 10^8 deg-K value is conservative. It is interesting to see that the slope of the noise versus frequency data given by Maxwell and Stone is roughly consistent

with a $1/f^{1.6}$ variation, or one that does not depart excessively from the inverse square relation observed at high frequencies.

In any event, the sky noise at these low, and very low frequencies is unusually large. When it is expressed as an effective temperature of the radiation resistance the temperature becomes a weighting factor such that the latter resistance may become the dominant noise generating component among the various resistances which make up the total antenna resistance. This definitely is the case for an electrically-large antenna and it may be true for the electrically-small antenna where the radiation resistance is very small.

3.2 Bases For Comparing Performance of Receiving Antennas

The advantage of an electrically-small, superconducting transmitting antenna seems rather clear; it is a matter of efficiency. If the reduced bandwidth is acceptable, a superconducting, electrically-small transmitting antenna wastes much less r.f. energy than its non-superconducting, but otherwise identical, counterpart when it radiates a given amount of r.f. energy. It is true that there may be matching problems and it may not be useful to eliminate all the ohmic resistance; however, down to where the bandwidth limitation becomes important, it is always advantageous, efficiency-wise, to reduce ohmic resistance. (In this respect, it should be noted that a reduction in ohmic resistance can be accomplished without the use of superconductivity).

It is much more difficult to define the advantages of electrically-small superconducting receiving antennas. Efficiency does not really enter because of the negligible power levels employed. (An exception would be found in the case where a transmitting-receiving system is used to transmit energy rather than information. It is doubtful that electrically-small antennas would be used for that purpose unless superdirective arrays of such small elements were to be found feasible; lossless superdirective arrays are consistent with single frequency transmission).

One can present an argument which seems to support the assertion that the potential for miniaturization is an advantage. The argument is based upon

a comparative analysis of the signal to noise ratio of antennas, the latter ratio being a criterion because a receiving antenna is not critical, usually, when a large signal level is available at the receiving site. (Again, we must qualify the statement because sometimes the directivity or perhaps the precision of the radiation pattern of the receiving antenna is the key factor, regardless of the received power density.) Suppose a standard, non-superconducting, tuned, half-wave dipole is compared to an electrically-small, superconducting antenna. Both antennas are operating at the same frequency and are exposed to the same incident signal. It will be recalled that the electrically-small antenna (either the magnetic or the electric dipole) has essentially the same theoretical aperture as the much larger half-wave dipole; the difference is approximately 8-percent in favor of the half-wave dipole. Accordingly, the available signal power captured from the incident wave is nearly the same under ideal conditions. The available noise power for either the small or the large antenna is given by the product kTB where k is Boltzmann's constant, T is the absolute temperature associated with the radiation resistance, and B is the noise bandwidth of the system. The temperature T, which is the effective temperature of the space into which the antenna is "looking", is predominant in the larger antenna because the ohmic resistance is negligible in comparison to the radiation resistance; it is predominant in the small, superconducting antenna because the ohmic resistance is largely eliminated.

If we compare the antennas on the basis of equal bandwidth (the latter may well be determined by the receiver circuits) it is easy to see that the conventional signal to noise ratio is nearly the same for both. The use of superconductivity in the case of the electrically-small antenna allows one to approach the performance of the large half-wave, non-superconducting dipole in the limit; therefore, the large antenna can be replaced by the nearly-as-effective miniature antenna.

This is a plausible argument and there are no really serious theoretical flaws in it. The first qualification would enter in connection with the bandwidth, and this only because we are using the conventional definition of signal to noise ratio. The word "available" applied to signal power and noise power implies maximum power transfer; and, maximum power transfer, in turn, implies matching the receiver input

resistance to the radiation resistance of the antenna by the use of a resonant transformer at worst and a conjugate reactance at best. Thus, a superconducting, electrically-small receiving antenna, used under conditions of maximum power transfer would operate in a tuned circuit having an exceedingly high Q-factor. The circuit itself might then determine the bandwidth of the receiving system and this bandwidth might be much less than what is desired or what is useful.

If an antenna-receiver system is operated under conditions of maximum power transfer and if the bandwidth afforded by an electrically-small, superconducting antenna is satisfactory, then there is a practical detail which may make difficult the attainment of the theoretically feasible miniaturization. While it is not difficult to match the 73 Ohm resistance of the half-wave dipole, one would be hard put to find a receiver with an input resistance as low as the radiation resistance of the miniature antenna. While this is not the fault of the antenna, it may as well be so long as such receivers are not available. If electrically-small, tuned, superconducting magnetic dipoles are used, the practical difficulty can be avoided by using the shunt connection to the antenna. This will magnify the radiation resistance by the square of the tuned circuit Q-factor. Normally, the resulting equivalent parallel resistance will not be greater than what is commonly available in input impedances of receivers. If an electrically-small, superconducting, electric dipole is used then an additional resonant transformer is required to permit the same effect. The transformer, which amounts to a coil, must be superconducting if the extreme impedance ratio is to be realized.

Quite often a low impedance antenna is connected directly to a high impedance receiver and no attempt is made to match impedances. This is true especially at the low frequency end of the radio frequency spectrum where antennas are electrically-small by necessity. In such an application the antenna does not limit the bandwidth, even if it is a superconducting antenna. The large series resistance of the circuit does not permit a high Q-factor even though a series conjugate reactance is inserted. Obviously, one detects the open circuit voltage of the antenna and the presence of the antenna impedance is not important to the signal voltage developed at the antenna terminals if the receiver input impedance approaches an infinite value.

But then, why eliminate the ohmic resistance of the antenna? A non-superconducting, electrically-small antenna should be as effective as a superconducting, electrically-small antenna if the antenna series impedance is ineffective.

The question is very pertinent and it is not easy to justify the use of superconductivity here. It can be justified but the advantage is marginal when presently available receivers are to be used. The advantage again lies in the possibility of miniaturization, and impressive miniaturization is predicted for an ideal receiver. Unfortunately, the degree of miniaturization which is possible falls off rapidly as the receiver departs from the ideal. The basis for comparison of receiving antennas used in this mode hinges on a new definition of signal to noise ratio. The discussion is lengthy and involved; consequently, we reserve the topic for the section of this report where it becomes part of an analysis of the particular mode of antenna operation. The reader is referred to sections 4.3 and 4.4 for a complete discussion.

3.3 A Figure Of Merit For Receiving Antennas

If a single figure of merit were used to rate receiving antennas, it probably would be derived from some sort of a signal to noise ratio because this latter would give an indication of antenna performance in areas of marginal signal intensity. The major function of superconductivity in receiving antenna applications is to provide for the possibility of miniaturization. Miniaturization is not meaningful or useful if the signal to noise ratio cannot be maintained. Superconductivity, by providing an increased available signal power in matched electrically-small antennas and reduced thermal noise in matched and unmatched electrically-small antennas, permits effective miniaturization.

It is suggested that the sensitivity per unit bandwidth of the antenna be used as a major criterion for the theoretical comparison of two antennas. These numbers can be calculated for the several modes of electrically-small antenna operation from the various expressions derived in section 4.

SECTION IV

MINIATURIZATION OF SUPERCONDUCTING ANTENNAS

4.1 General Discussion of Concepts and Criteria

It would be well at the outset to define certain terms and concepts. By miniature antenna we mean an antenna that is electrically-small (Its physical dimensions are a small fraction of the longest wavelength radiated or absorbed by the antenna), to the degree that it is much smally physically than the conventional antenna it might replace. Thus, we would not be impressed by a microwave antenna which fits into a one inch cube, but a highly efficient 500 kHz antenna which fits into the same volume would receive our instant attention.

The word antenna is used in its most general sense. It may refer to a simple dipole, to an array, or to some other configuration or collection of elementary or complex radiators. The context in which the word is used or some specific comment will identify the item when that is necessary. An elementary radiator is any element which approximates the incremental electric or magnetic dipole.

There is another aspect to miniaturization where the establishment of some sort of quantitative criteria is prerequisite to really meaningful discussion. It is not enough for an antenna to be miniature, miniaturization must be achieved without the undue sacrifice of other important antenna characteristics.

We offer two criteria which may be used to quantify miniaturization. The first is applicable to the miniaturization of existing design. The second anticipates possible innovation.

Case I. - There exists a conventional antenna which serves its purpose well in some system except that something could be gained by reducing the physical size

of the antenna. No other change in specifications can be tolerated. Assuming that the larger antenna represents the optimum design, true miniaturization will have been achieved if a new, smaller and lighter (at least, no heavier) superconducting antenna system provides the identical performance.

Note the implication in the qualifying words lighter and system. The system is inclusive of whatever refrigeration equipment may be needed; it will do little good, in general, if the combination of miniature antenna plus refrigeration system occupies more space and weighs more than the larger antenna.

Case II. - Take any arbitrary superconducting antenna, no matter what its specifications may be, and build around it a useful receiving (or transmitting) system so that some significant requirement within the limits of the arbitrary specifications is satisfied. If this same requirement cannot be filled by a smaller and lighter system utilizing a non-superconducting antenna, then effective miniaturization will have been accomplished.

Again, the size and weight of any refrigeration must be taken into account. Further, the reader must not overlook the latitude implied by the latter use of the word system.

Let us now survey the miniaturization of (superconducting) antennas in the light of the introductory discussion to see if the boundaries of performance and size can be closed at this time.

Are there any fundamental limitations to electrically-small superconducting antennas? One can anticipate, with Wheeler [32], a difficulty in matching the extremely low radiation resistance of the elementary radiator because this resistance decreases more rapidly with miniaturization than does the ohmic resistance. This difficulty may be more severe in the case of a transmitting antenna than a receiving antenna. It is hoped that the matching problem can be resolved for signal levels and frequencies where superconductivity can eliminate, effectively, the ohmic resistance of both the radiator and the matching network.

The bandwidth problem is more serious. Antenna and matching

network reactance in conjunction with the very small radiation resistance remaining after ohmic resistance is eliminated make for extremely high-Q circuits. When tuning is necessary, a simple tuned circuit may well produce a bandwidth that is insufficient for useful information rates.

Directivity seems limited to values near that for the elementary radiator. This is not a bad thing; when one wants an omnidirectional or a quasi-omnidirectional antenna, one avoids large directivity values. However, there are also applications for highly directive antennas. The problem is this: A highly directive antenna invariably consists of a spacial superposition (lumped or continuous) of elementary radiators. The interference pattern of the composition gives the desired effect. Practical highly directive antennas thus occupy spacial dimensions measuring many wavelengths. Now the miniaturization of the <u>elements</u> in the distribution is possible without sacrifice of directivity, but the second step, miniaturization of the distribution by reduction of element spacing, obviously destroys the usual interference relations. The conventional high directivity or high gain antenna cannot be miniaturized in the latter sense.

Subsequent sections will examine the technological aspects of miniaturization achieved through superconductivity to see whether or not effective miniaturization is possible, in view of material, insulation, and refrigeration requirements. At the present time it is not possible to escape the need for a cryostat (with the possible exception of an application of natural cooling); the cryostat and its back-up facilities must be sufficiently adequate for useful missions. However, we need not limit our thinking on this account because the future may bring new superconducting materials operating at higher temperatures. Cryostat or Dewar requirements are greatly eased as temperatures increase. Finally, the most speculative thinking envisions room temperature superconductors, an achievement that would eliminate the cryostat entirely.

In one sense it is not entirely fair to view the matching problem as a limitation of ideal electrically-small antennas. This fault is more properly assigned to the transmitter or to the receiver, and it becomes another case where associated technology is lagging. Nevertheless, even though superb antennas were feasible,

there is no advantage if they cannot be used with existing equipment. A study of what is possible now is given as much attention in the following work as a study of what may be possible in the future.

The bandwidth problem is more severe in the case of transmitting antennas than in the case of receiving antennas. It is not always necessary to match receiving antennas, especially in the frequency ranges where electrically-small antennas are most useful. Consequently, the matching problem largely disappears and the antenna no longer limits the bandwidth of the system. These aspects are explored in the sequel.

Directivity may well remain a fundamental limitation of the electrically-small antenna. The only hope is the possibility for a superdirective antenna. The prospects are very dim but they are examined in a following sub-section.

4.2 Incremental Electric and Magnetic Dipoles

There are two types of elementary antennas, the incremental electric dipole and the incremental magnetic dipole. Most electrically-small antennas tend to become one or the other in the limit as the size is reduced for a given frequency or as the frequency is reduced for a given size.

The electric dipole is familiar to everyone. The small, or incremental electric dipole is merely a shortened version; often, in order to maintain a uniform current over the length of the antenna (as is assumed in conventional analysis) the small antenna is end-loaded by flat plates or large balls. In the limit, the electric dipole resembles a parallel plate capacitor.

The incremental magnetic dipole is a small loop antenna; it may be a simple one-turn loop or a multi-turn coil or solenoid, just so long as its dimensions are small compared with the free space wavelength corresponding to the frequency of the exciting current. The extended length of the wire in a multi-turn coil also should be small in comparison with the free space wavelength. Normally, the incremental magnetic dipole is pictured as a tiny, circular, one-turn loop.

It is appropriate that we display, for reference in this report, the basic electromagnetic expressions describing the circuit properties and the radiation fields for these incremental radiators. The derivations are found in standard textbooks [33, 34]. In the field expressions, time is suppressed and the conventional spherical polar coordinate variables are employed. The radius vector from the orgin to the point in question has the magnitude r. The polar angle is θ and the asimuthal angle is Φ. The electric dipole lies along the polar axis and the magnetic dipole has its surface normal to the polar axis. In every case the current I is uniform in all parts of the conductor, and the symbol designates the amplitude of the time-harmonic current of angular frequency $\omega = 2\pi f$.

The incremental electric dipole is characterized by the following expressions:

$$E_\theta = \frac{I\,l}{4\pi\epsilon_o}\left[\frac{j\omega}{c^2 r} + \frac{1}{c\,r^2} + \frac{1}{j\omega r^3}\right]\sin\theta\; e^{-j\beta r} \tag{4-1}$$

$$E_r = \frac{I\,l}{2\pi\epsilon_o}\left[\frac{1}{c\,r^2} + \frac{1}{j\omega r^3}\right]\cos\theta\; e^{-j\beta r} \tag{4-2}$$

$$H_\Phi = \frac{I\,l}{4\pi}\left[\frac{j\omega}{c\,r} + \frac{1}{r^2}\right]\sin\theta\; e^{-j\beta r} \tag{4-3}$$

Radiation resistance:

$$R_a = \frac{\eta}{6\pi}(\beta\,l)^2 \tag{4-4}$$

Series capacitance:

$$C = \frac{\pi\epsilon_o l}{2\log_\epsilon (l/a)} \tag{4-5}$$

where a is the wire radius, l is the total length of the element, $\beta = 2\pi/\lambda = \omega/c$, and $\eta = \sqrt{\mu_o/\epsilon_o}$.

The incremental magnetic dipole is characterized by the following expressions:

Antenna fields:

$$H_\theta = - \frac{\beta^2 I S}{4 \pi r} \left[1 + \frac{1}{j\beta r} - \frac{1}{\beta^2 r^2} \right] \sin\theta \; e^{-j\beta r} \tag{4-6}$$

$$H_r = \frac{j \beta I S}{2 \pi r^2} \left[1 + \frac{1}{j\beta r} \right] \cos\theta \; e^{-j\beta r} \tag{4-7}$$

$$E_\phi = \frac{\eta \beta^2 I S}{4 \pi r} \left[1 + \frac{1}{j\beta r} \right] \sin\theta \; e^{-j\beta r} \tag{4-8}$$

Radiation resistance:

$$R_a = \frac{\eta}{6\pi} \beta^4 S^2 \tag{4-9}$$

Series inductance (one-turn circular loop):

$$L = \mu_0 b \log_e (b/a) \tag{4-10}$$

where b is the loop radius, a is the wire radius, and S is the loop area. If there are N turns, I is the current per turn multiplied by N; also tobe multiplied by N^2 are R_a and L. (All units are MKS)

All the expressions listed above are derived without taking into consideration the ohmic resistance of the conductors; consequently, the conventional application of superconductivity will permit, at best, the realization of whatever may be predicted by these formulas. No change is brought about at any step in the derivation if superconductivity is assumed from the start.

The size-independent aperture, or cross-section, is readily predicted from these expressions. In the case of the incremental electric dipole, the only electric field component which survives over remote distances is the inverse r part of E_θ. The available signal power from the dipole becomes

$$P = \frac{E_\theta^2 \, l^2}{8 \, R_a} \qquad (4\text{-}11)$$

where $E_\theta l$ is the amplitude of the induced harmonic voltage. The delivered power absorbed from the plane wave is

$$P = \frac{E_\theta^2 \, A}{2 \, \eta} \qquad (4\text{-}12)$$

where A is the defined aperture, or cross-section, of the antenna which intercepts the average power density $E_\theta^2 / 2\eta$ present in the wavefront, When the expressions are equated and R_a is substituted, it readily is seen that

$$A = (3/8\,\pi)\,\lambda^2 \qquad (4\text{-}13)$$

or

$$A = 0.119\,\lambda^2 \qquad (4\text{-}14)$$

which is independent of antenna size, the latter having cancelled out in the process of derivation.

In the case of the incremental magnetic dipole, the amplitude of the induced harmonic voltage for a one-turn loop is given by [34]

$$V = E_\Phi \, \beta \, S \qquad (4\text{-}15)$$

where E_Φ is the inverse r part of that expression. Thus, the available signal power from the magnetic dipole is

$$P = \frac{E_\theta^2 \, A}{2 \, \eta} \qquad (4\text{-}17)$$

to the load. Equating and substituting the expression for loop R_a yields

$$A = (3/8\pi)\lambda^2 \tag{4-18}$$

as before for the electric dipole. The conclusion is the same.

In consulting the literature [35] for the aperture of a half wave linear electric dipole, one finds

$$A = 0.13\lambda^2 \tag{4-19}$$

Futher, it is found that the half wave linear electric dipole is only slightly more directional than the incremental dipole (either type), specifically, the respective directivities are 1.64 and 1.5

These results are thought provoking. Why should we use a large dipole when practically the same power can be intercepted by one of vanishing size and when there is no important loss of directivity? The question is particularly significant in the case of low frequency antennas where a half wavelength may be very long physically.

4.3 Untuned Incremental Dipoles

Is a receiving antenna really necessary? The question has some significance because it is possible to manufacture an argument which seems to support the assertion that a receiving antenna is required only to compensate for the shortcomings of the receiver. The reasoning goes somewhat as follows: Imagine that we have a perfect receiver. The specifications for such an item would include adjustable but unlimited gain, a noise factor of unity (or zero decibels), and, in this case, an infinite input impedance over the frequency range to be received. The bandwidth of the receiver is arbitrary; it may be adjusted to any reasonable value.

If we had such a receiver, there would be no minimum value for the voltage which could be detected; it would be necessary only to increase the gain until the amplified voltage or its analog appeared at the output. The usable gain is

unlimited because the receiver itself does not generate noise. Naturally, the input voltage cannot be zero, that is meaningless, but it can be arbitrarily close to zero.

Now, let the receiver input voltage be derived from an incremental electric dipole connected to the input terminals. The rms voltage induced in the dipole would be $\bar{E}\, l$ where l is the length of the dipole and $\overline{E}$ is the rms electric field intensity of the incident electromagnetic wave. This field may contain sky noise or atmospheric noise as well as a signal. In addition, the antenna will develop a thermal noise voltage due to its ohmic resistance and effective series ohmic resistances due to coupling. In all, a total rms voltage V will be developed by the antenna. At this point of the discussion we do not care if the voltage consists of a signal or noise or a combination thereof, the perfect receiver will amplify whatever it "finds" at the antenna terminals.

How much of this total developed voltage will appear at the antenna terminals? All of it will appear there because the receiver has an infinite input impedance over the bandwidth. No current flows; therefore, no voltage drop takes place in the finite antenna resistance and reactance; the receiver sees the open circuit potential of the source. It is immaterial that the incremental dipole has reactance and it is not necessary to tune it out.

Finally, we can let the length l of the incremental dipole become arbitrarily small because the input voltage can become arbitrarily small; and the antenna disappears into the input terminals of the receiver.

It is to be understood that we are not advocating the feasibility of the perfect receiver. This is merely a thought experiment taking place at a theoretical limit, an experiment which may help us gain a true picture of the function of a receiving antenna. It is important always to differentiate between theoretical limits and practical limits so that we do not give up a possibility by mistaking a technological limitation for a theoretical limitation. Gains in technology often permit a closer approach to what is possible theoretically.

Two aspects require some additional scrutiny; these are antenna reactance and antenna noise. As the incremental electric dipole is shortened, its

series capacitive reactance increases; however, it can become arbitrarily large because we have postulated an infinite input impedance for the receiver. The portion of antenna noise about which we are concerned is that due to the ohmic resistance of the antenna conductor. The sky or atmospheric noise is mixed with the signal, both arrive as electromagnetic waves and cannot be distinguished from each other except by techniques which do not enter the domain of our investigation.

There are a number of ways of interpreting the behavior of the conductor ohmic resistance thermal noise voltage as the incremental electric dipole length is decreased. On the one hand we might observe that the ohmic resistance of the conductor varies directly as the length of the dipole. Consequently, the thermal noise voltage will vary as the square root of the dipole length and it will not fall off as fast as the induced voltage which varies as the first power of the dipole length. Some of the induced voltage consists of sky noise which is calculated by assigning a temperature to the radiation resistance and then utilizing the usual thermal noise expression. This noise will fall off at the same rate as the signal as the length is decreased because the radiation resistance varies as the square of the length. Thus, the noise contribution from the ohmic resistance of the conductor will eventually overcome the signal as the dipole length is reduced.

In another interpretation, one might wish to reduce the diameter of the dipole simultaneously as the length is reduc~~ed so that~~ a thin antenna is maintained. If this is done, it is easily seen that the ohmic resistance decreases even more slowly and it may even remain constant as the length of the incremental dipole is decreased. Consequently, it appears that the conductor ohmic resistance will finally determine the sensitivity of the arbitrarily short electric dipole unless the diameter can be increased as the length is decreased.

There are two ways in which the noise contribution of the ohmic resistance can be nearly eliminated. Superconductivity can eliminate the ohmic resistance (however, the a.c. behavior of superconductivity must be examined carefully to make sure that significant noise is not contributed thereby); reduction of conductor temperature can eliminate noise equally well because both temperature and resistance appear as first power variables in the radical which yields the rms noise voltage,

namely $\sqrt{4kTRB}$. Neither the theory of superconductivity nor the theory of extremely low temperatures (quantum statistical mechanics) is sufficiently advanced to tell us which method will bring us closer to zero noise.

Now, let us put these thoughts together. The perfect receiver will allow us to reduce the length of the incremental electric dipole connected to its terminals to the degree that the dipole becomes of vanishing size, provided either that the dipole is manufactured from a superconducting material or that the temperature of the dipole can be reduced to zero. But, as this size limit is approached, how is one able to distinguish between the antenna and the actual terminals of the receiver? These terminals must have some portion that is exposed to free space; that is, there must be some interface between the receiver and the external world. In order to obtain a single result, we conjecture that it is easier to obtain zero ohmic resistance via superconductivity than it is to approach arbitrarily close to absolute zero.

The conclusion that then seems to follow is this: When everything else is perfected in the receiver, in order to obtain a zero decibel noise figure, it is still necessary to eliminate the ohmic resistance in the connector that forms the interface between the receiver and the radiation field; otherwise the Johnson noise generated in this connector prevents attainment of perfection. But what is this vanishingly small resistanceless (ohmic) connector other than the ultimate electrically-small, superconducting antenna?

It is in this sense that one can say that a perfect receiver does not require an antenna because (of the somewhat contradictory fact) an antenna already is an inseparable part of every receiver; further, if the receiver is perfect, this antenna is electrically-small and it is superconducting.

This same argument, in so far as it holds water, probably represents the ultimate defense for the assertion that superconductivity will improve the electrically-small receiving antenna.

Another fact that emerged in the discussion is the broadband behavior of this untuned incremental dipole. The perfect receiver possesses an infinite input impedance over the arbitrary passband; it will not load the antenna; therefore, the

open circuit potential of the antenna appears over the entire passband of the receiver. It is not necessary to resonate the antenna reactance.

This almost incidental fact forms the topic of this section. It is interesting to see that, theoretically at least, there is some basis for a broadband electrically-small antenna that also will be useful for receiving marginal signal levels. The next step is to determine how closely these ideal conditions can be approached in practice.

Let us begin by examining the concept of the perfect receiver. Unlimited gain cannot be achieved but this specification can be approached more closely than the others. Through careful shielding and decoupling, and by translating the frequency band, one can almost always obtain much more gain than is actual usable. The receiver bandwidth will be decreased as the gain is increased because amplifiers have defined gain-bandwidth products; however, present gain-bandwidth products are very large and ever increasing. A usable bandwidth certainly would be available for any usable gain that can be achieved.

The specification, infinite input impedance, probably comes closer to encountering a fundamental limitation than the others. The usual limitation is the shunt capacitance developed across the input terminals. Some capacitance seems unavoidable so long as we have two conductors separated by an insulating medium. The presence of a leakage resistance also is unavoidable; however, present devices have such high values, 10^{12} Ohms and higher, that the shunt capacitance represents the major departure from the ideal. A reduction of shunt capacitance to 0.1 pF would be a major accomplishment at the present time, but the reactance of even this low value is down to 10^{12} Ohms at only 1.6 Hz.

A zero decibel noise figure specification represents another limit which can only be approached; however, modern amplifiers can come surprisingly close. For instance, some parametric amplifiers have noise figures of only a fraction of a decibel for frequencies below 500 MHz. The unfortunate thing is that extremely low noise figures are not always useful in receiver front ends at the lower frequencies because of the large amount of sky noise. This latter fact may dilute

the motivation for seeking marginal improvement in amplifier noise figure in that frequency range.

We assume that the practical receiver can be shielded sufficiently well and that frequency translation can be used, if necessary, to prevent significant feedback from reaching the input terminals of the receiver. The most obvious source of external feedback is the output port of the receiver. This problem can be controlled by frequency translation.

A more serious difficulty is found in the geometry of the receiver. The input terminals must be located somewhere in the surrounding shell. Consequently, the receiver shell unavoidably is in the near field of the electrically-small antenna; the shell becomes a parasitic element or a ground plane, depending on the wavelength. As a parasitic element or a ground plane it will modify the antenna pattern but, even more important, it will introduce noise, the amount depending upon the degree of coupling, electric and/or magnetic. It is likely that a superconducting material would have to be used for the shell. Quite possibly, a physically small preamplifier could be located at the terminals and the main receiver elsewhere.

The bandwidth of the untuned electric dipole probably will be larger than the bandwidth of the receiver if very high gain is to be used; and, if an unusually sensitive antenna is employed then the gain will be high whenever the antenna is used to its maximum effectiveness. For purposes of analysis, one could pick an arbitrary number, say 10-percent, for the bandwidth.

What degree of performance is available when the untuned incremental electric dipole is used with an imperfect receiver which, nevertheless, approaches the perfect receiver as a limit? Will the use of superconductivity improve the performance as it would in the case of a perfect receiver?

The problem, as analyzed below, deals with one imperfection at a time. First, consider the effect of non-infinite input impedance where the remainder of the receiver is perfect. This finite input impedance can be considered in parallel with an infinite input impedance, perfect receiver. We will lump the finite impedance with the antenna and calculate the performance of the composite circuit taken as a

whole.

Figure 1 illustrates the latter.

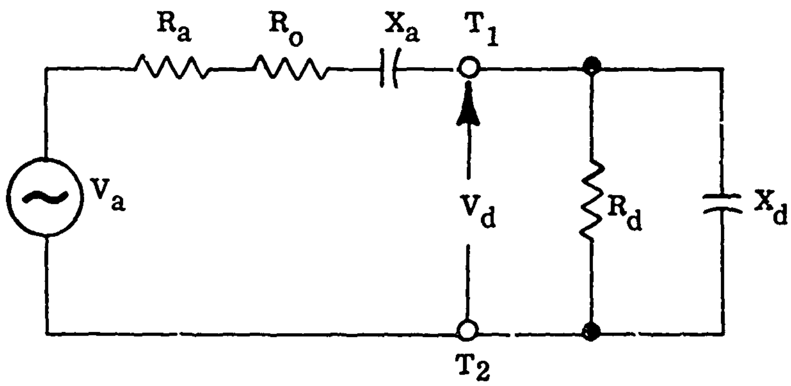

Figure 1. Signal Equivalent Circuit of Antenna and Receiver Input.

The terminals $T_1 - T_2$ represent the antenna-receiver interface. Define $R = R_a + R_o$ where R_a is the radiation resistance and R_o is the ohmic resistance. Reactance X_a is that of the antenna; it is very much greater than R for the electrically-small antenna. Receiver input resistance and shunt capacitive reactance are R_d and X_d. Voltage V_a is that induced in the antenna by the electromagnetic fields; voltage V_d is that found at the receiver terminals.

It can be shown that the terminal voltage is given by the expression

$$V_d \approx V_a \left[(1/Q_{ad}^2) + (1 + D)^2 \right]^{-1/2} \tag{4-20}$$

where $Q_{ad} = R_d / X_a$ and $D = X_a / X_d$.

The indicated approximation is very good; we neglect antenna resistance R in sums where other terms are greater by factors of 10^4 and more. When this expression is used in calculations, it should be remembered that the loss tangent of the capacitor dielectric is important. The effective parallel resistance of the capacitor is to be included in the calculation of R_d. One should be suspicious of R_d values which exceed the magnitude of X_d by a factor of more than 10^4.

The noise equivalent circuit applicable to the problem is shown in figure 2.

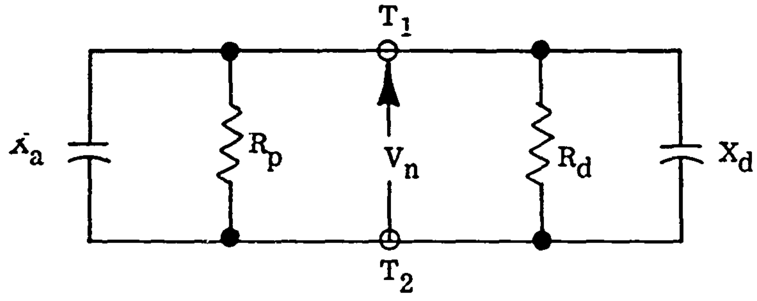

Figure 2. Noise equivalent circuit.

Because $X_a >> R$, it is possible to accurately represent the series antenna impedance by the parallel equivalent circuit drawn to the left of the terminals. Resistance $R_p = X_a^2 / R$. It also is convenient to define a dimensionless quantity $Q_a = X_a / R$.

Subsequent calculations will be based upon the assumption that all these components are at the same temperature. The more complicated case can be handled as shown in van der Ziel [29].

Figure 3 now shows the series equivalent circuit for the circuit of figure 2. The noise resistance R_n is given in terms of the other quantities defined above.

$$R_n \approx R \left[\frac{1 + \frac{Q_a}{Q_{ad}}}{(1 + D)^2 + \frac{1}{Q_{ad}^2}} \right] \tag{4-21}$$

The indicated approximation neglects the addition of the reciprocal of antenna Q_a to unity in one step of the derivation.

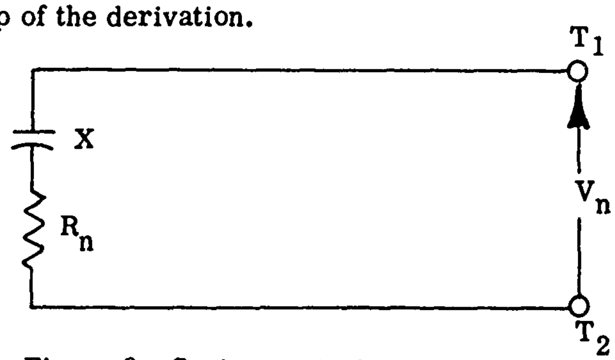

Figure 3. Series equivalent noise circuit

The rms thermal noise voltage is given by $\bar{V}_n = \sqrt{4kTBR_n}$. Reactance X is not essential to this calculation because the circuit works into an infinite impedance.

An indication of the ultimate performance of the system is given by an expression which is a normalized ratio of signal voltage to noise voltage. The basis of normalization is the signal voltage to noise voltage ratio available to the perfect receiver. Designate the normalized ratio by K such that

$$K = (V_d / \sqrt{4kTBR_n}) / (V_a / \sqrt{4kTBR}) \qquad (4\text{-}22)$$

Then K is a number which shows how closely a receiver, imperfect by reason only of finite input impedance, approaches the performance of the perfect receiver using the same incremental dipole. Factor K becomes unity at this limit. Substitution yields

$$K^2 = 1/(1 + (Q_a / Q_{ad})) \qquad (4\text{-}23)$$

For a given antenna Q_a is fixed at a given frequency and it does not vary rapidly with frequency so that all depends on the quantity Q_{ad} . It can be seen that good performance requires an extremely large input resistance for the imperfect receiver under discussion. This latter resistance should be greater than the numerical antenna reactance by a factor which in turn is at least an order of magnitude greater than the ratio of antenna reactance to total antenna series resistance.

The result is most interesting because it is independent of the input shunt reactance of the receiver. When the shunt reactance is relatively low, the antenna reactance and the shunt reactance is relatively low, the antenna reactance and the receiver shunt reactance form a capacitance divider for the signal voltage; but, as can be seen, the square root of the noise resistance is reduced in a matching proportion. Thus, within the limitation discussed earlier, we can tolerate input shunt reactance and still obtain good performance. The fallacy is, of course, that any practical receiver will be imperfect beyond the finite input impedance; there will be receiver noise other than that generated thermally in the input impedance, and this noise will not decrease when the signal is reduced by the capacitance divider.

Now let us pick two electrically-small, electric dipoles so that the previous expressions can be interpreted numerically in connection with something that can be visualized physically. Both antennas are balanced, center-fed dipoles, and it is assumed that the receiver input is balanced. The operating frequency is approximately 16 MHz. The important facts and calculations are summarized below. The ohmic resistance calculation is based on copper.

Table I. Summary Of Antenna Dimensions And Parameters.

	Antenna 1	Antenna 2
Full length in cm.	2	20
Diameter in mm.	1	1
C_a in pF	0.1	0.5
X_a in Ohms	10^5	2×10^4
R_a in Ohms	10^{-3}	0.1
R_o in Ohms	10^{-2}	0.1
Q_a	10^7	10^5
R_p in Ohms	10^{12}	2×10^9

At this operating frequency we might hope to find a receiver with a shunt input reactance of 10^5 Ohms and a shunt input resistance of 10^{10} Ohms. Substituting appropriate values in expression (4-23) yields $K^2 = 0.01$ for Antenna 1 and $K^2 = 0.8$ for Antenna 2. These numbers are indicative of signal to noise ratio deterioration due to finite receiver input impedance. In both cases, the antennas would perform equally well if a perfect receiver were available; however, the shorter antenna suffers a severe loss in performance even with this almost unrealizable input impedance specification. The performance of the longer antenna might be

deemed acceptable.

Suppose superconductivity were employed to reduce or eliminate R_0; it would be possible to take advantage of the noise reduction, but a significant gain would occur only where Q_a / Q_{ad} is less than unity. Thus, the normalization factor increases as R is reduced, but so does Q_a.

We have analyzed the effect of one imperfection, now let us examine the limitations introduced by another. The noise generated by the active devices in the receiver will ultimately limit the usefulness of the antenna-input impedance sensitivity predicted by the numerator of expression (4-22). The signal level simply must be raised until it overcomes this noise; increased input impedance or superconductivity cannot help.

The principle of duality automatically causes one to search for the zero input impedance, perfect receiver with a broadband performance brought about by the elimination of the necessity for tuning. Such a dual condition exists. The above receiver would be driven by an incremental magnetic dipole which, in the limit, would consist of a short circuit across the input terminals of the perfect zero input impedance receiver.

At first glance it appears as though the incremental magnetic dipole could not be a frequency independent current source because of the unavoidable series inductive reactance; however, while the inductive reactance increases directly as the frequency, the induced voltage does likewise (see expression (4-15)) so that the short circuit current of a resistanceless incremental magnetic dipole is independent of frequency. Of course, the closed circuit consisting of the magnetic dipole and the receiver input cannot be entirely resistanceless, otherwise only a persistent current could flow if the circuit were superconducting.

Another possibility for a broadband untuned, electrically-small antenna is the case where a magnetic dipole is used with a high input impedance receiver. The results here are nearly the same as for the incremental electric dipole connected to the high input impedance receiver and expression (4-20) holds, provided inductive reactance X_a is small compared with shunt capacitive input

reactance X_d .

In conclusion, it is possible to have effective untuned, electrically-small receiving antennas which will provide broadband operation. A very high input impedance, low noise receiver must be used. Superconductivity will provide only marginal benefits except in the case of the ultra-low noise receiver where it is important to take advantage of every possibility for achieving noise reduction.

4.4 Electrically-Small, Tuned Magnetic Dipoles At The Low Frequency Limit

The previous section discussed interesting possibilities for electrically-small antennas when these were used in conjunction with noise free receivers having extremely large values for input resistance R_d, or slightly noisy receivers having extremely large values for input impedance Z_d. When receiver noise approaches zero, it becomes advantageous to use superconducting, untuned, electrically-small receiving antennas.

Since few practical receivers possess the high input resistance or impedance demanded by the previous theory, one searches for ways of accommodating the typical receiver. One clue lies in the fact that the magnitude of R_d is a relative thing. It must be large when compared with X_a, the antenna reactance. But there are two kinds of reactance, one being the anti-reactance of the other; accordingly, it seems logical that a reduction of antenna reactance could extend the range of expression (4-22) to lower R_d values. In other words, we may "tune" antenna reactance X_a with a conjugate reactance. Because total antenna resistance R for an electrically-small antenna is still very much less than any R_d value found in typical high or even moderate input impedance receivers, it will be true that $V_d = V_a$ and $R_n = R$, and that the reduction of R by superconductivity may prove to be beneficial.

This possibility is gained at some cost because reactance and anti-reactance both are frequency dependent but in inverse ways so that complete cancellation of X_a takes place only at one frequency; significant cancellation is achieved only in a relatively narrow frequency interval. A study of expressions (4-21) through

(4-23) shows that the latter frequency interval is governed by a deterioration factor

$$D_n = X_{at} / \sqrt{RR_d} \tag{4-24}$$

where X_{at} is the algebraic sum of the antenna reactance and the conjugate reactance. The quantity K^2, which is related to a type of signal to noise factor for the antenna plus the immediate receiver input circuit, approaches its highest value when D_n approaches zero; and, a 3 db loss is sustained when D_n becomes unity. This means that the effective bandwidth for the ideal signal to noise ratio is given by

$$B_n = f_o / Q_n \tag{4-25}$$

where f_o is the center frequency where the reactances cancel and factor Q_n is defined by the ratio $X_a / \sqrt{RR_d}$. Reactance X_a is the antenna reactance (only) at the center frequency.

If R_d is relatively large, the bandwidth restriction need not be too severe. Antennas 1 and 2 of Table I in the previous section, if tuned, would have, respectively, useful low noise bandwidths, B_n, of 5 kHz and 80 kHz with a center frequency of 16 MHz when used with a receiver having an input impedance consisting of 10^5 Ohms in parallel with as much as 10 pF. Bandwidth B_n is not to be confused with the much larger signal bandwidth of the low-Q series resonant circuit.

Although either tuned incremental electric dipoles or incremental magnetic dipoles can be used in the manner discussed above, there would seem to be some advantage in using the magnetic dipole because inductance is required in either case. Moreover, magnetic dipoles are commonly used for electrically-small antennas at low frequencies; the published performance of these would furnish a basis for comparison of any results of analysis obtained here.

A basis for comparison of receiving antennas seems to be the antenna sensitivity; however, we do not wish to employ the usual concept of sensitivity which is based on available signal and available noise because this implies maximum power trnasfer. Our antennas will operate into high impedances so that the terminal voltages

for both signal and noise will be the open circuit values. With this in mind, we define a signal to noise ratio which is the square of the antenna open circuit signal voltage divided by the square of the rms open circuit noise voltage developed by the antenna resistance R.

Thus,

$$SNR = V_a^2 / 4kT_n B_n R \tag{4-26}$$

where T_n is the effective noise temperature (necessary to account for cases where the components of R are at different temperatures, see van der Ziel [29] p. 15) and B_n is the noise bandwidth discussed previously. The voltage sensitivity of the antenna is defined here as the V_a corresponding to SRN = 1; however, sometimes it is convenient to express this sensitivity in terms of the electric field intensity in the incident wave which will induce the latter value of V_a.

When two antennas are compared, the total noise environment of the antennas must be considered; this normally includes the "sky" noise, or the "temperature" of its radiation resistance (which has no direct, or first order, connection with the actual temperature of the physical antenna), noise power due to the ambient temperature of the various ohmic resistances associated with the physical antenna and its ancillary structures, and noise power due to the input resistance of the detector or amplifier. The question is whether or not one antenna is theoretically capable of performing as well as another at the same locale. The value depends also on bandwidth, hence, we must specify equal bandwidths for the two antennas. It should go without saying that the band-center frequencies are the same.

A theoretical analysis performed at the low frequency limit avoids many problems that could cause controversy if high frequencies were compared. Primarily, there would be certain embarrassing questions relating to impedance levels. The treatment of directivity, which is so easily obtained in conventional antenna at high frequency, would necessitate numerous definitions and qualifications. But if we go to the low frequency limit, our superconducting antenna is on an equal footing with the conventional antenna because <u>all</u> antennas become electrically-small

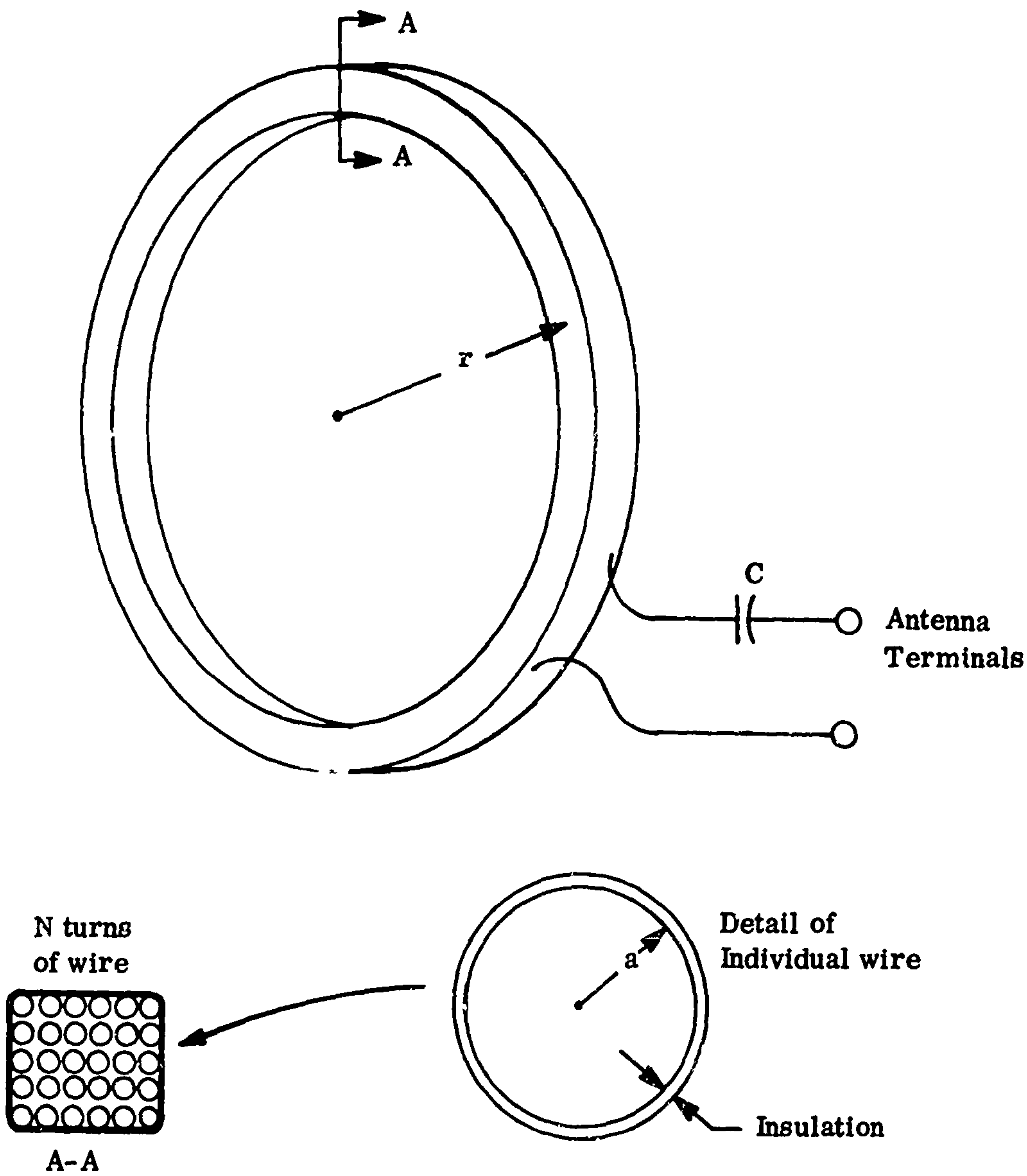

Figure 4. Details of the Magnetic Dipole Loop Antenna Considered in the Text.

antennas at low enough frequencies; the role of superconductivity will stand out with unmistakable clarity, if there is truly an advantage.

First, one must recognize that the electrically-very small antenna has a negligibly small radiation resistance and, even in the case of a superconducting, electrically-very small antenna this is not the resistance which determines the Q. It seems beyond belief that a 1 Hz antenna approximately 1 meter in extension would have a Q of the order of 10^{24}, but such is predicted by the Q expression of section 5.1. It is intuitive that the Q will be much lower in magnitude. This feeling is supported by solid experimental evidence, such as, known losses caused by dielectric materials and distributed lossy material in the near electrical vicinity, that is, the near field. (It would be difficult to escape from the near electrical vicinity of an antenna operating at a wavelength of 3×10^{8} meters).

We must assume at the outset that the major resistance, even of the superconducting antenna, is effective ohmic resistance introduced by coupling into surrounding dissipative media or by actual shunt resistance caused by unavoidable dielectric loss.

There is a flaw in this argument which must be rationalized before we proceed further. While it is true that the radiation resistance is negligible in value from the standpoint of the signal circuit, it is not negligible from the standpoint of noise. At the low frequencies considered, the natural noise fields found on the surface of this planet are unusually intense [31]. If this noise is converted into an effective temperature, the radiation resistance, while negligible for Q calculations, becomes the predominant value for the noise calculation. Because we are interested in the possibility of extending our results to higher frequencies, we will for the moment take the liberty of defining the natural noise field as a signal. This at least is valid in instances where measurement of the natural noise field is the objective and in some instances where deep strata communications is contemplated. Accordingly, we will assume the radiation resistance to be at the earth temperature of approximately 290 deg-K.

Let us then deal with an electrically-small loop antenna. As in explained in section 6.7, a loop antenna suffers considerably less from the effects of surrounding dissipative media than a short electric monopole or dipole; therefore a loop antenna

seems a logical choice for an extremely low frequency, electrically-small antenna. A series of expressions applicable to the loop antenna of figure 4 are developed in the following paragraphs.

The area of the loop is

$$S = \pi r^2 \qquad (4\text{-}27)$$

where r is the mean radius of the circular loop.

The resistance of the loop is

$$R = 2\pi r N P \qquad (4\text{-}28)$$

where N is the number of turns and P is defined as the effective resistance per unit length of wire. It is important to remember that the radiation resistance will be negligible under the conditions stated previously. Resistance R contains not only the ohmic resistance of the wire but also any effective series resistance due to dielectric losses and coupling losses.

The inductance of the loop is given approximately by expression (4-10) if we multiply it by N^2 and interpret dimension a as the radius of the bundle of N wires. To simplify matters, let us assume that the loop radius to bundle radius ratio always will be near 10. This will allow the $\log_e (b/a)$ factor to be replaced approximately by 2.5, then

$$L_a = 2.5\,\mu_o\, r\, N^2 \qquad (4\text{-}29)$$

All units are MKS and $\mu_o = 4\pi \times 10^{-7}$.

The Q-factor of the loop antenna, provided we are not near the self resonance frequency, is given by

$$Q_a = (5 \times 10^{-7}\omega\, N)/P \qquad (4\text{-}30)$$

where ω is the angular frequency.

The defined factor Q_n, can be calculated from the expression

$$Q_n = 2.5\, \mu_o\, \omega \sqrt{\frac{r\, N^3}{2\pi\, P\, R_d}} \qquad (4\text{-}31)$$

The low noise bandwidth may be calculated by using this Q_n in expression (4-25) provided that Q_n is greater than 10. If the factor is much smaller than 10, the bandwidth B_n must be interpreted differently. For large values of Q_n

$$B_n = \frac{1}{2.5\, \mu_o} \sqrt{\frac{P\, R_d}{2\pi\, r\, N^3}} \qquad (4\text{-}32)$$

When an electromagnetic wave with electric field intensity E is incident upon the loop, oriented for maximum signal, the open circuit voltage induced in the loop is given by the expression (see expression (4-15))

$$V_a = E\,(2\pi^2/\lambda)\, r^2 N \qquad (4\text{-}33)$$

where λ is the signal wavelength.

We are now in a position to calculate the signal to noise ratio as defined by expression (4-26). The calculation should be modified, however, because Q_n will be much less than 10 at the low frequency limit. Because of this, the low noise bandwidth will be quite wide and the noise bandwidth will be determined by the receiver bandwidth, even in the case of a superconducting loop. It is more meaningful to calculate the SNR factor on a per Hertz basis, thus

$$\text{SNR - Hz} = (E^2\, \pi^3\, r^3\, N) / (2k\, \lambda^2\, T_n P) \qquad (4\text{-}34)$$

where Boltzmann's constant $k = 1.38 \times 10^{-23}$ Joule per deg-K and T_n is the previously discussed effective noise temperature of P.

Suppose two antennas at two temperatures are to be compared as follows: The SNR-Hz is to be the same for given E, λ, and N.

Then

$$(r_2 / r_1) = (T_2 P_2 / T_1 P_1)^{1/3} \tag{4-35}$$

Since P is a function of temperature, it would appear that the radius of the low temperature antenna could be reduced drastically. For example, if the resistance per unit length could be reduced by a factor of 100 when the effective temperature is reduced to 4 deg-K from 300 deg-K, the radius of the loop could be reduced by a factor of 20.

Unfortunately, this calculation carries the implicit assumption that there are noiseless active elements in the initial stages of the high gain receiver. The SNR-Hz is retained by the small loop but its open circuit voltage has decreased by a factor of 400; which presents no problem to the assumed receiver, but which may prevent useful application where the receiver has a fixed device noise. Each receiver application thus becomes a separate problem where an optimization is performed on expressions (4-33) and (4-35) relative to the measured noise figure of the receiver. The value of expression (4-32) should always be checked and the device noise contribution must be separated from the noise figure.

4.5 Performance Of Electrically-Small, Tuned Magnetic Dipoles In A 300 deg-K Radiation Environment

Section 5 will discuss the Q-factors and the expected bandwidths of electrically-small, superconducting transmitting antennas. The natural Q-factors, those determined by the radiation resistance alone, will be high; however, the actual Q-factors will more likely be determined by various loss mechanisms that are difficult or impossible to avoid (see sections 6.5 through 6.9).

We speak of magnetic dipoles in the title; actually, any tuned electrically-small superconducting antenna will be a composite of a low-loss capacitor and a low-loss inductor. Such elements are the limiting form of the electric dipole and the magnetic dipole respectively. Most of the electromagnetic energy will be stored in the reactive fields; the antenna function is provided via the small amount

of energy which leaks into the radiation field. The element from which the preponderance of this energy leaks is, usually by design, the radiating element. In fact, if energy leaks equally well from both, it can be arranged by proper mechanical orientation that the combination forms a circularly polarized antenna. Since a coil will be present in any case, it may be well in most instances to utilize the magnetic dipole as the radiator because it does not couple as strongly into surrounding lossy dielectric media (see section 6.7) and because the problem of end loading is circumvented.

Electrically-small transmitting antennas probably will be tuned if they are to be driven by existing generators. If they are superconducting antennas, the efficiency of the antenna will be much greater than that for the normal type. Whether or not the efficiency of the system is greater than that of the normal system depends upon the coupling network and the transmitter. Since it is not possible now to obtain extremely low series impedance and series resistance transmitters with the exception of the charged capacitor used as a transmitter, see section 7.1, the usual high impedance transmitter can drive the tuned superconducting antenna in the shunt mode.

Let us make an order of magnitude calculation on an antenna with the critical dimension d approximately 1-percent of the wavelength. Dimension d would be the total length of an incremental electric dipole or the radius of an incremental magnetic dipole. Section 4.2 provides us with theoretical values of the radiation resistances and expression (5-2) gives us a sufficiently good estimate to the theoretical Q-factor. For the electric dipole the radiation resistance is approximately 0.1 Ohm and the theoretical Q-factor is approximately 10^5. This makes for an antenna reactance of approximately 10^4 Ohms. The actual Q-factor, which takes into account additional components of series resistance due to dielectric losses and the like, may well be 10^4, assuming an order of magnitude increase in the resistance; consequently, the tuned antenna presents a shunt mode resistance of approximately 10^8 Ohms to the transmitter. It will be difficult for the typical transmitter to deliver appreciable power to such a large resistance, although the operation will be efficient from an energy point of view because the transmitter has a much lower internal resistance.

The radiation efficiency of the system will be approximately 10-percent,

mainly because of lossy distributed material in the near field of the antenna; however, this material and the corresponding loss exist even for the normal antenna so that the superconducting antenna still is more efficient than the latter. The loss which is eliminated is that due to the ohmic resistance of the conductor making up the antenna and the tuning coil.

Neither the series mode nor the shunt mode drive is appropriate for this tuned antenna. The tuning coil should be tapped at the proper position so that the transmitter sees a lower resistive load. The tuning coil must be superconducting and the system must be driven as a balanced load so that no large r.f. currents flow in non-superconducting materials. A monopole using the earth or even high conductivity materials for a ground plane cannot take advantage of superconductive operation because of the large losses encountered by the antenna currents which must flow in the reflecting plane. This, probably, is the reason for the limited success in the Moore and Travers [18] experiment.

If the antenna is a magnetic dipole instead of an electric dipole, the radiation resistance is approximately 3×10^{-3} Ohms for a 1-turn loop, the Q-factor still is of the order of 10^5 so that the antenna reactance is approximately 300 Ohms. It should be possible to hold the Q-factor above 10^4 in practice so that the effective shunt mode resistance presented to a source is approximately 3×10^6 Ohms. This is lower than that of the tuned electric dipole but it is still too high for realistic practical operation. Again the loop may be tapped or a capacitance divider can be used to lower the resistance. Balanced operation is vital, as before.

Within the bandwidth of operation, both antenna types will show a system efficiency of approximately 10-percent provided the transmitter impedance is relatively low. This efficiency is at least 100-percent better than that which could be achieved by normal electrically-small antennas of the same size. The efficiency will not approach the much higher value obtainable with electrically-large, non-superconducting antennas. If the electrically-small, superconducting antenna is driven from a low impedance generator, the bandwidth in which the relatively good efficiency is obtained will be narrow, only 100 parts per million in the cases studied. The actual operating bandwidth will be greater, provided the generator

can supply the increased current without saturating (at higher frequencies for the electric dipole and lower frequencies for the magnetic dipole), but the efficiency will deteriorate rapidly because of the internal losses of t' generator.

These are order of magnitude calculations for a d/λ ratio of 1-percent. They are apparently on the pessimistic side because some VLF electrically-small antennas with a d/λ ratio of 1-percent are reported to exhibit antenna efficiencies near 10-percent. (See Jasik, [36] section 19.) In view of these well documented measurements, we may hope to construct superconducting antennas having this d/λ ratio with system efficiencies of 50-percent and higher because the reported losses can be eliminated by superconductivity, at least in the case of similar antennas scaled to the megahertz region.

Thus, in regard to transmitting antennas, there is no doubt that superconductivity will permit very efficient electrically-small transmitting antennas, but at some cost in bandwidth. Whether or not high power electrically-small, superconducting, transmitting antennas are feasible depends upon the outcome of further r.f. tests on Type II materials (see section 2.3). It must be emphasized that miniaturization is the only direct advantage. The superconducting, electrically-small transmitting antenna competes only with the non-superconducting, electrically-small transmitting antenna. It cannot match the performance of the non-superconducting, half-wave transmitting dipole or quarter-wave transmitting monopole.

Now let us examine the anticipated performance of electrically-small, tuned, receiving antennas operating into a matched load. This condition exists when maximum power transfer is the object, or when reflections are to be avoided in electrically-long transmission lines connect antenna and receiver, or when the amplifier into which the antenna works is sensitive to variations in VSWR. Once more the reader is reminded that the anticipated region of application lies below the mid-UHF range where antennas already are physically-small.

When receiving antennas are matched for maximum power transfer, the best basis for comparison is the relative sensitivity as determined by the conventional signal to noise ratio. It was shown in section 3 that for a given bandwidth, the signal to noise ratio (available signal and available noise) of the lossless

electrically-small dipole can approach, but not exceed, that of the half wave dipole. This statement holds true both for electrically-small electric dipoles and electrically-small magnetic dipoles. In dealing with tuned superconducting elements, the logical choice is the magnetic dipole because a tuning coil will be required in any case; it may as well be used as the antenna.

The analysis follows that of section 4.4 to a certain extent. For an electrically-small magnetic dipole the radiation resistance will be so small that it usually can be neglected in comparison with the other effective series resistance which remains after the coil becomes superconducting. A problem would arise when the effective temperature of the radiation resistance would be so high that the noise contribution would be great despite the small resistance value. We escape this difficulty by limiting the analysis to a 300 deg-K environment; meaning that the results hold for antennas used above the surface of the earth and operating at frequencies above some 120 MHz, or for antennas used below the surface of the earth, in deep strata communications [37], and operating at any desired frequency. Receiving antennas using frequencies much below 120 MHz above the surface of the earth are subject to sky noise [30] and a special analysis must be made for each application to be sure that the radiation resistance temperature does not become too heavy a weighting factor; it is not always possible to take advantage of the inherent signal to thermal noise ratio of antennas used in that frequency range.

For the matched antenna, the expression for signal to noise ratio resembles expression (4-26). The available signal is $V_a^2 / 4R$ where V_a and R are as defined in that section. The available noise is $kT_n B_n$ where T_n is the effective noise temperature, determined as indicated in section 4.4, and B_n is the noise bandwidth. This latter value may be determined by the receiver, or it may be determined by the loaded antenna. In the latter instance, the value will be the center frequency in Hertz divided by one-half the Q-factor of the antenna. (one-half, because the antenna drives a matched load).

Thus

$$B_n = P/(5\pi \times 10^{-7} N) \qquad (4\text{-}36)$$

If we solve for the signal to noise ratio on a per Hertz basis, the resulting expression will be the same as expression (4-34). The numerical results will be different because effective temperature T_n will reflect the heavy contribution by the matched load. The temperature will be much higher unless the load resistance can somehow be cooled. The signal to noise ratio for the matched antenna is calculated by dividing expression (4-34) by expression (4-36) and the sensitivity of the magnetic dipole can then be calculated by finding the electric field intensity E for which the signal to noise ratio is unity. One should always check the noise bandwidth predicted by expression (4-36) to be sure that it is not ridiculously small for the intended application.

To summarize: In comparing antenna performance, one should begin by finding all components of the antenna series effective resistance and the load resistance and the temperatures associated with these components, being especially careful to determine the "sky temperature" if the frequency is below 120 MHz. An effective noise temperature now is calculated after the procedure outlined in van der Ziel [29] p. 15. The bandwidth is computed from expression (4-36) and the signal to noise ratio is obtained using expression (4-34). The superconducting, electrically-small magnetic dipole will invariably have a better signal to noise ratio than an identical antenna that is non-superconducting, but the bandwidth will be smaller. It will be up to the user to decide if the increase in sensitivity is worth the loss in bandwidth. In no case will the signal to noise ratio for a bandwidth equal to that of the superconducting, electrically-small antenna ever exceed that of a non-superconducting, half wave dipole. The electrically-small, superconducting antenna competes only with electrically-small, non-superconducting antennas.

4.6 Broadbanding Low Frequency, Tuned Magnetic Dipoles

It is apparent from the previous sections that the major benefit is gained by cooling the receiving antenna. The advantages of superconductivity cannot be fully realized unless an ultra low noise receiver is used with the antenna. Further, when superconductivity is employed, the bandwidth of the antenna may narrow excessively

with certain types of antenna loading.

In view of the above, superconductivity may become useful in some applications by providing a means for controlling bandwidth. The temperature may be reduced to eliminate as much thermal noise as is possible and to eliminate conductor ohmic resistance. Then, ohmic resistance can be added deliberately to reach the value that gives exactly the desired bandwidth. The secret is to contain the added resistance in the low temperature bath so that it contributes a negligible amount of thermal noise. In some instances the increase in conductivity with decreasing temperature of non-superconducting materials is sufficient to reach the desired condition. For example, the ohmic resistance of copper reduces sufficiently at liquid nitrogen temperatures (by a factor of 7, [38] that the skin effect, or a.c., resistance is reduced by a factor of 2.5. The a.c. resistance would be reduced by a factor of 30 at liquid helium temperatures, corresponding to a resistivity reduction factor of 900. It would be possible to attain Q-factors of 10^4 without resorting to superconducting materials.

This broadbanding technique would not be applicable to transmitting antennas because we are not interested in reducing thermal noise. Even a cold resistance will convert electrical energy into heat. The antenna would be more efficient but at the expense of a considerable boil-off of cryogenic fluids.

4.7 Signal To Noise Ratio-Bandwidth Product

We have seen in the previous sections that a signal to noise ratio factor is useful in comparing the performance of electrically-small antennas, in particular, superconducting, electrically-small antennas and non-superconducting, electrically-small antennas. It is patent that as the temperature of an antenna is reduced, the signal to noise ratio increases and the bandwidth decreases. The latter is true because the bandwidth is a function of the antenna resistance and the resistance, in turn, is a function of the temperature.

It would be unfortunate, indeed, if the gain in signal to noise ratio were exactly offset by a loss in bandwidth. Then, the whole concept of improving

antennas by cooling would become meaningless because it already is possible to improve signal to noise ratio by narrowing the bandwidth with a bandpass filter. (Granted, the bandpass filter would need to be lossless in order to produce an equivalent effect). It becomes important to demonstrate that a genuine improvement accrues from cooling.

Assume that the antenna is exposed to a constant radiation field intensity; then, the signal to noise ratio can be expressed as

$$SNR = A/(TRB) \tag{4-37}$$

where A is a constant, T is the absolute temperature, R is the antenna effective resistance, and B is the effective bandwidth of the antenna. (Our concern is groundless if the antenna does not determine the bandwidth of the system; expression (4-34) clearly predicts an improvement with decreasing temperature).

The antenna effective resistance R is a function of the temperature, especially when ohmic resistance is present. If we begin our comparison at normal ambient temperatures, the conductor ohmic resistance presents the greatest contribution in the case of an electrically-small antenna. Reference shows [38] that it is not bad to use a T^2 dependence for the volume resistivity of conductors over the wide temperature range envisioned here. This dependence changes at temperatures near absolute zero. In the latter range it tends toward a constant. The resistance of the r.f. conductor will not exhibit a T^2 variation because of skin effect. As the temperature is decreased the skin thickness decreases because the resistivity decreases; accordingly, the r.f. conductor resistance does not decrease as rapidly with temperature as does the resistivity. It turns out that the resistance varies as the square root of the resistivity. We will make the reasonable assumption that the conductor r.f. ohmic resistance varies as the absolute temperature T.

$$R = C_1 T \tag{4-38}$$

The bandwidth of the antenna varies directly as the resistance in the

case of a matched antenna and it varies as the square root of the resistance in the case of broadband operation. Since we would be interested mainly in the former case where antenna bandwidth is likely to determine the system bandwidth, we will assume the direct variation. But resistance varies directly with temperature, therefore

$$B = C_2 T \tag{4-39}$$

Substitution in expression (4-37) yields

$$SNR = C_3 / T^3 \tag{4-40}$$

The most revealing relation is the product of the signal to noise ratio and the bandwidth. Multiplying expression (4-39) into expression (4-40) yields

$$(SNR)(B) = C_4 / T^2 \tag{4-41}$$

and it is seen that the SNR-bandwidth product improves rapidly as the antenna temperature is reduced from the normal ambient. It is important to realize that this condition holds for the electrically-small antenna where the conductor ohmic resistance is much larger than the radiation resistance and the other incidental resistances at the usual ambient temperatures. Little would be gained by reducing the temperature of an electrically-large antenna because here the radiation resistance is large compared to the ohmic resistance.

Even when the antenna is working under the broadband type of operation discussed in a previous section, the SNR-bandwidth product varies as the inverse square of the absolute temperature. Temperature reduction always seems to improve the inherent performance of electrically-small antennas. Great improvement is obtained even at liquid nitrogen temperatures; thus, at 77 deg-K the SNR-bandwidth ratio becomes 15 times as large as it is for the same antenna operated at 300 deg-K.

We do not wish to mislead the reader with impressive numbers. These figures represent potential gains in performance; the degree to which they are

realizable depends upon the type of receiver used, the frequency of operation, the information rate requirements, the noise environment, and perhaps several other factors. But some gain is almost always possible and some trade-off with antenna size seems assured especially where a low noise receiver is able to utilize the inherent sensitivity of the antenna and where the sky noise is not excessive. As was mentioned before, sweeping generalizations are not possible; each application must be analyzed separately in the light of available equipment and the operating conditions. In addition, it should be recalled that the gains are not unlimited. It was shown in section 3 that the performance of the electrically-small antenna cannot exceed that of the half-wave dipole operating into the same bandwidth at the same temperature. Expression (4-41) loses its accuracy at temperatures in the vicinity of 10 deg-K or at any temperature where the conductor ohmic resistance reduces to the same order of magnitude as the radiation resistance or the other incidental effective series resistances associated with an antenna. Moreover, it loses significance when the effective noise temperature of the antenna differs appreciably from the ambient temperature of the antenna. In general, it is advantageous to go all the way to the superconducting mode unless the effective temperature of the radiation resistance (the sky temperature) becomes exceedingly high, provided, of course, that the bandwidth still is useful and that the receiver noise figure is sufficiently low.

4.8 Superdirectivity

Thoughts of superdirectivity enter almost automatically into the study of the electrically-small antenna because superdirectivity seems to offer the only hope for building a direc electrically-small antenna. Otherwise an electrically-small antenna seems doomed to take on the form of an incremental electric or magnetic dipole with the attendant quasi-omnidirectional characteristic.

Naturally, one can build a linear, areal, or spatial array of miniature elements but they will have to be spaced by inter-element distances that approximate a quarter-wavelength; the array itself will not be electrically-small; that is, superconductivity may be able to permit miniaturization of the radiating elements but it will not permit the miniaturization of free space. Occasionally there may be an

advantage in generating a highly directive interference pattern by using two widely spaced miniature elements instead of two widely spaced electrically-large elements, but generally we wish the entire antenna system to be electrically-small.

If the elements of the array are spaced at intervals that are electrically-small, then it is possible in theory to drive the elements with predictable current magnitudes and phases that produce a highly directive pattern; this is the superdirective array. Harvey [15] gives a sufficiently complete set of references to permit a survey of the present status of the theory. The outlook is not good because the resulting array possesses an exceedingly high Q-factor and it is necessary to maintain a fantastically high precision in mechanical alignment and in the current magnitude and phase values.

The reasoning that would tend to connect superconducting, electrically-small antennas with superdirectivity is somewhat vague and intuitive, but there is a correlation that should not be entirely ignored. The first item of correlation is the high Q-factor of the superdirective array: A superconducting element can have an exceedingly high Q-factor; its value increases as the element decreases in electrical size. The second item of correlation is the need for precise mechanical orientation: Superconducting elements can become vanishingly small; accordingly, the mechanical resolution can increase, almost without limit. This is an indirect benefit of miniaturization; the efficient miniature antenna element approaches a truly point source of dipole radiation. There is another subtle aspect to the possibility of element miniaturization. The minimum inter-element spacing is determined by the physical size of the element itself. If large elements are spaced too closely, there is a great deal of unwanted electrostatic or capacitive coupling between the individual elements. Consider the following analogy: The capacitive coupling between 5-inch diameter spheres spaced on 1-foot centers is considerably greater than the capacitive coupling between 5-mil spheres spaced on 1-foot centers. A third item of correlation is the need for precise current magnitude and phase relations between the elements: Superconducting, tuned, electrically-small loops that are spaced at electrically-small distances form coupled circuits with interesting current relationships. See section 6.2 for a discussion of the theory. For an

electrically-small spacing the current ratio for two such loops is unity. It would be interesting to discover the current distribution of a multi-loop array with close spaced elements, one being the driven element. It seems intuitive that the current distribution would settle into a very well defined and stable configuration. When used as a receiving array, for example, this exact configuration might be very sensitive to the direction of approach of the exciting wave.

Thus, there is sufficient correlation to make for an interesting theoretical problem and sufficient justification for an interesting experiment. It is unlikely that the theoretical approach will lead to a positive or negative conclusion; it would have done so years ago what with all the attention focused on the matter. It is suggested that a collection of really miniature, tuned, superconducting, magnetic dipoles, such as are discussed in section 8, be placed in an r.f. radiation transparent cryostat, such as is described in that same section, and a simple test be conducted in search of directive properties.

One closing thought is a reiteration of the observation that the transmission of electromagnetic energy at a single frequency is consistent with the concept of a superconducting-superdirective antenna. Schelkunoff and Friis ([34] p. 198) are particularly lucid on this point.

SECTION V

PHYSICAL SHAPE FACTORS

5.1 Q-Factor Versus Antenna Shape

In the case of the elementary radiator, whether it be an electric or a magnetic dipole, the radiation resistance will be exceedingly small if the element is electrically-very small. The reactance will be high; consequently, a tuned radiator will display very high system Q, leading to bandwidth problems. Wheeler [32] has included this as one of the fundamental limitations of small antennas. Since the Q-factor of an antenna is some function of antenna shape, it becomes important to discover the range of the variation.

All the investigation into this factor has led to the tentative conclusion that the adjustment of Q by variation in physical shape is limited in range to one or two orders of magnitude. This, or course, excludes the obvious method of increasing Q by adding ohmic resistance; we are interested in the interplay between the reactance X_a of the electrically-small antenna and the radiation resistance R_a of the same. The Q-factor in question is defined by

$$Q_a = X_a \,/\, R_a \tag{5-1}$$

Both X_a and R_a depend upon the physical dimensions of the electrically-small antenna, and by idealizing the problem to the point where the antenna, either an electric dipole or a magnetic dipole, conforms to the dimensions of a cube of side d, it can be shown that the antenna Q_a-factor becomes (see, for example, Wheeler above)

$$Q_a = (3/4\pi^2)\,(\lambda\,/\,d)^3 \tag{5-2}$$

where λ is the wavelength in the same units as the length d.

Naturally, the actual antenna need not have the form of a cube, but it seems intuitive that the orthogonal dimensions will give a geometric mean value near that of d when the latter dimensions all differ. Some variations can be expected but it seems doubtful that a high order of magnitude adjustment can be achieved. Smith and Johnson [39] have demonstrated this fact in connection with electrically-small, (non-superconducting) broadcast antennas. By top loading the antennas, the radiation resistance could be doubled while at the same time the reactance was halved. This is a four-to-one variation in Q introduced by varying the physical shape.

It is well known that short, fat electric dipoles have smaller Q-factors than thin dipoles of the same length. Also, when the magnetic dipole is flattened into a plane loop, the loop wire radius in the log (b/ a) factor of expression (4-10) predicts a lower Q for a thick wire.

Another possibility that comes to mind is the use of a self-resonant helix or spiral. The Q-factor of a self-resonant coil is quite low and it is possible, but far from certain, that this condition will extend to the superconducting helix or spiral where the inductive reactance is resonated by the reactance of the distributed capacitance of the coil. There is some plausible logic for this conjecture. The collection of turn-inductances and interturn-capacitances might be viewed as a series of coupled circuits where the individual resonances spread into a band.

From the description of multiple-tuned VLF antennas given by Martin and Carter in Jasik [36] one gathers that relatively low Q-factors are obtained in these physically-large and elaborate, but electrically-small antennas. A given antenna will have an effective length corresponding to 1-percent of the operating wavelength but the reactance divided by the radiation resistance is only 1500, approximately. This is in contrast to the 10^5 value predicted by expression (5-2). The actual Q-factor, due mainly to antenna ohmic resistance is more like 150. Still, this gives roughly a 10-percent radiation efficiency, which includes the loss in the matching elements, and a reasonably good bandwidth. To be fair, one should take into account that the antenna system is spread over a volume which would make the geometric

mean length d equal to 5-percent of the wavelength. Substituting this in expression (5-2) yields a predicted Q-factor of approximately 1000. In this sense the theoretical expression may be very realistic. Nevertheless, this type of antenna may be a model worth additional study because much of the large volume is needed to reduce ohmic resistance, which can be eliminated (in a scaled version) by superconductivity. (Note: This type of antenna uses a ground plane. A scaled version, for higher frequencies, would need to be converted into a balanced system.)

The idea, of course, is to make the Q_a value as small as possible so that the antenna will have a usable bandwidth. To place this problem in a proper perspective, let us identify the antenna bandwidth. We have seen elsewhere in this study that it is not realistic to expect the actual Q of the antenna to be equal to Q_a because there are unavoidable losses in nearly every practical case, losses that enter as effective series ohmic resistance even though the ohmic resistance of the metallic conductor forming the antenna vanishes due to superconductivity. If the value X_a / R_o is defined as Q_o where R_o is the total effective series ohmic resistance of the antenna, then the actual Q-factor of the antenna is found from the relation

$$1/Q = 1/Q_a + 1/Q_o \tag{5-3}$$

so that the bandwidth $B = f/Q$ is given by

$$B = f/Q_a + f/Q_o \tag{5-4}$$

where f is the frequency at the band center.

An increased bandwidth, while necessary for a higher information rate, is not an unmixed blessing. It permits more noise power to enter the communication channel. In this sense, the reduction in the effective Q of the antenna by the ohmic resistance and the resultant bandwidth increase is gained at not too great a loss in signal to noise ratio because the temperature of much of the ohmic resistance is likely to be that of liquid helium while the temperature of the radiation resistance may be greater, perhaps much greater, than 300 degrees Kelvin. In other

words, sometimes the effect of physical shape upon the Q of a superconducting antenna is of somewhat less concern for the practical superconducting antenna than for the theoretical superconducting antenna.

5.2 Expected Bandwidth Range

Expression (5-2) will give a theoretical value for the Q-factor of a tuned, electrically-small antenna if the critical dimension d is known. A geometric mean of the three orthogonal dimensions will at least give an order of magnitude result. Note that though this expression arises from ideal incremental dipole theory, its derivation is not contingent upon the use superconductivity. Superconductivity merely permits a practical approach to the ideal condition.

Suppose that dimension d is approximately 1-percent of the wavelength; then the Q-factor will be of the order of 10^5 and the tuned antenna will exhibit a bandwidth of the crder of 10 parts per 10^6. It is doubtful, however, that such a high Q-factor will be attained by any matched, tuned, superconducting, electrically-small antenna exposed to the radiation field. With all the loss mechanisms operating, it may be difficult to reach a Q-factor of 10^4 and a corresponding bandwidth of 100 parts per 10^6. It definitely will be possible to attain a Q-factor of 1000 and even 1500; such values have been measured in our laboratory for 20 MHz magnetic dipoles in foamed polystyrene cryostats, unshielded and exposed to the radiation environment, and only at liquid nitrogen temperature.

Unless the Q-factor can be reduced by more than an order of magnitude through physical shape adjustment, the bandwidth probably will be determined by the effective series ohmic resistance caused by losses in the immediate and nearby dielectric materials, nearby parasitic structures, and distributed lossy media in the near field. If the antenna is a monopole and the earth is used as a ground plane, the source of the low Q-factor is rather obvious. A superconducting ground plane is required if the monopole is to behave in accordance with ideal theory. Naturally, the induced ohmic resistance will not cause heat to be developed in the cryostat and will therefore not interfere with the thermal efficiency of the latter.

(Dielectric losses internal to the cryostat are an exception). The losses in the external media heat the external media.

Thus, the expected bandwidth for superconducting, tuned, electrically-small antennas that are less than 1-percent of the wavelength long will be of the order of 100 parts per 10^6. Perhaps the physical shape of the radiator can be adjusted to improve the bandwidth by reducing Q through reactance decrease rather than resistance increase; however, this must remain as an item for further investigation.

The theoretical calculation of antenna Q-factor for arbitrary shapes quickly leads to intractable mathematics. To make matters worse, not enough is known about current distributions on various shapes of superconductors. It is recommended that theoretical efforts in this direction be abandoned in favor of an experimental approach where a series of intuitively promising shapes are tested.

SECTION VI

LONG RANGE MAGNETIC COUPLING

6.1 Introduction

In coupled circuit theory the mutual inductance between two isolated and electrostatically shielded coils is the circuit concept used to explain the transfer of energy from one circuit to the other. The transformer is the most common practical application of the physical phenomenon which is involved, and the same phenomenon forms part of the experimental basis for Faraday's law of induction.

There always is some physical separation between the two coils, even in the case of bifilar wound transformers, and the mutual inductance is a function of this separation. It becomes interesting to apply the magnetically coupled circuit concept to coils that are separated by distances that are meaningful in wireless communications. What usually is neglected in the conventional circuit analysis is the time retardation in the response of the second circuit to a signal applied to the first circuit, a retardation brought about by the finite speed of propagation of the magnetic field. Such a delay is present in every case; however, it may be insignificant. The criterion probably would be the smallness of the ratio of circuit spacing to the free space wavelength corresponding to the highest important frequency in the signal.

In what follows, the time retardation is included as part of the analysis of the coupled circuits. The two coils are really electrically-small, magnetic dipole antennas and it develops that superconductivity is required for a practical approach to the theoretical predictions. In this light every transformer, every magnetically coupled coil pair, can be viewed as a transmitting-receiving antenna application. The most interesting cases are those where the coils are physically small in comparison with their separation.

6.2 Ideal Theory Of Magnetic Coupling

Consider a situation where a generator drives a small superconducting loop. Another superconducting loop of the same size (for mathematical simplicity only) is located r units away. The transmitting loop may not require tuning but the receiving loop is tuned. The two identical loops may be considered coupled inductors with M as the mutual inductance.

We may represent the magnetically coupled inductors by an equivalent "tee" circuit as shown in figure 5.

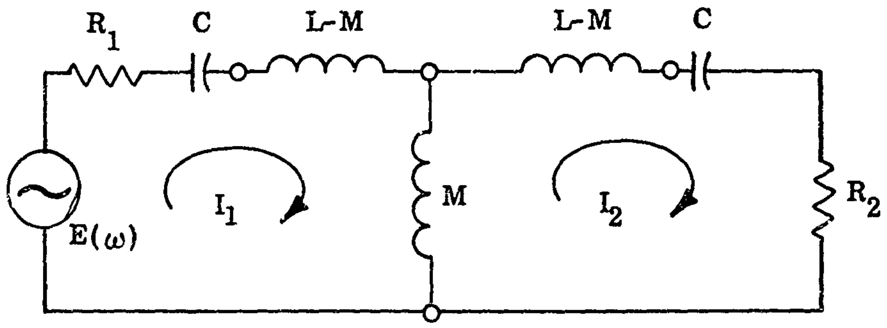

Figure 5. Equivalent circuit for the coupled loops.

In a long range experiment, one where the coil separation is much greater than the physical dimensions of the coils, the coupling will be weak, that is, M will be very small. Assume that at angular frequency ω_0 the following expression holds

$$\omega_0^2 \, C \, (L-M) = 1 \tag{6-1}$$

Then the net reactance of an inductance (L-M) in series with a capacitor C becomes zero and the first equivalent circuit reduces to the one illustrated in figure 6.

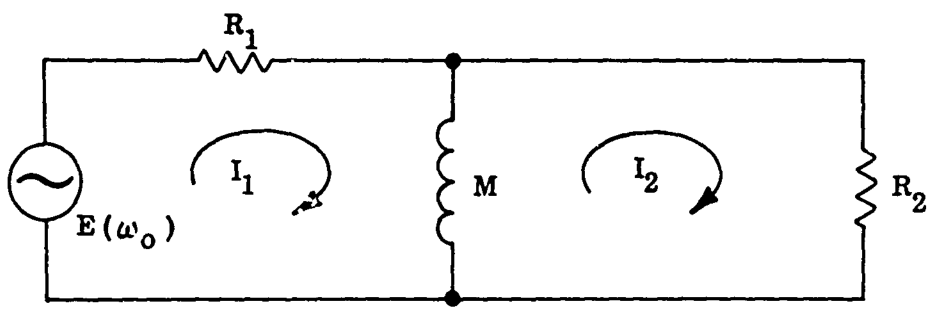

Figure 6. Equivalent circuit at frequency f_o.

Resistance R_2 ideally is the radiation resistance of the small loop (the ohmic resistance being zero due to superconductivity). Resistance R_1 is the radiation resistance of the first loop plus the generator resistance, which may be ohmic resistance. Thus R_1 will be very much larger than R_2 (by orders of magnitude) and R_1 will be very much larger than the magnitude of ω_o M because of the long range assumption; consequently, the voltage source and R_1 may be replaced by a current source. Figure 7 illustrates this final reduction.

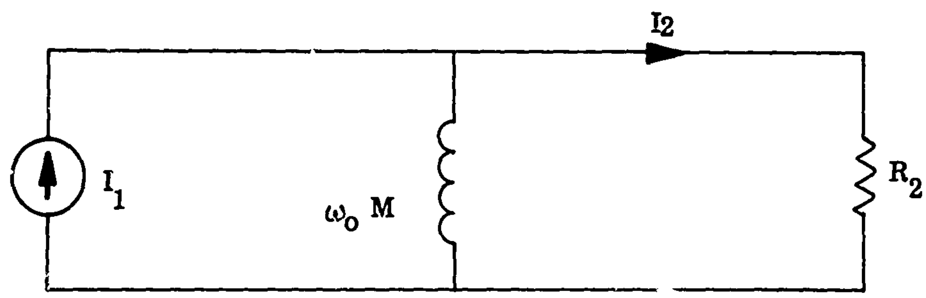

Figure 7. Reduced equivalent circuit at frequency f_o.

Next we must derive an expression for M. Once this is done, it will be possible to predict current I_2 in the receiving loop.

The mutual inductance is defined by the expression

$$M = (N_2 \Phi_{12}) / I_1 \qquad (6-2)$$

where N_2 = turns on coil 2, Φ_{12} = flux in coil 2 due to current I_1 in coil 1.

But

$$\Phi_{12} = \mu_o H_{12} S \tag{6-3}$$

where S is the cross-sectional area of the receiving loop which is taken to be oriented normal to the H_{12} magnetic field intensity direction and μ_o is the absolute permeability.

Schelkunoff and Friis [34] show that the H_θ field of a small loop when transit delay is taken into account has a magnitude given by the expression

$$H_\theta = \frac{\beta^2 IS}{4\pi r}\left(1 + \frac{1}{j\beta r} - \frac{1}{\beta^2 r^2}\right)\sin\theta \tag{6-4}$$

where S = area of loop, $\beta = 2\pi/\lambda$ where λ is the free space wavelength of the signal, r is the radial distance from the loop to where the field is measured, and I is the current in the loop.

When the two loops are lying in the same plane $\theta = \pi/2$; hence, by substitution

$$M = \frac{\mu_o \beta^2 S^2}{4\pi r}\left(1 + \frac{1}{j\beta r} - \frac{1}{\beta^2 r^2}\right) \tag{6-5}$$

where we have recognized that $H_\theta = H_{12}$, $I = I_1$, and $N_2 = 1$ Turn.

The same authors also develop the radiation resistance of a small one-turn loop. The final expression is

$$R = (\eta/6\pi)/(\beta^2 S)^2 \tag{6-6}$$

where $\eta = \sqrt{\mu_o/\epsilon_o}$ and ϵ_o is the absolute permittivity. Resistance R_2 will be given by this expression.

Treating, in figure 7, the mutual inductance M and radiation resistance R_2 as a parallel circuit and using the current division formula, one finds that

$$I_2 / I_1 = (j\,\omega_o M) / (R_2 + j\,\omega_o M) \qquad (6\text{-}7)$$

There are various possibilities.

First, if $R_2 \ll \omega_o M$ then $I_2 = I_1$ (6-8)

Second, if $R_2 = \omega_o M$ then $I_2 = I_1 / \sqrt{2}$ (6-9)

Third, if $R_2 \gg \omega_o M$ then $I_2 = (\omega_o M / R_2)\, I_1$ (6-10)

In connection with this third case, substitution yields

$$\omega_o M / R_2 = \frac{3}{4\pi} \frac{\lambda_o}{r} \left[1 + \frac{\lambda_o}{j 2 \pi r} - \frac{\lambda_o^2}{4 \pi^2 r^2} \right] \qquad (6\text{-}11)$$

and it is seen that the third condition, $R_2 \gg \omega_o M$, is satisfied for large values of r. In other words, the third condition applies to the case where the two loops are separated by a large radial distance. At these remote distances, the last two terms in the square brackets may be neglected, i.e., $\lambda_o \ll r$, and the expression becomes

$$\omega_o M / R_2 = (3/4\pi)\,(\lambda_o / r) \qquad (6\text{-}12)$$

Substitution in the third case now yields

$$I_2 = (3/4\pi)\,(\lambda_o / r)\, I_1 \qquad (6\text{-}13)$$

when the second loop is remote from the first. The interesting observation here is that the result is independent of the area of the loops.

The simple result, $I_2 / I_1 = 1$, which at first glance is relegated to the close range problem, has important implications which deserve further comment. With reference to figure 7, current I_1 is that of the transmitting loop antenna and I_2 is the receiving loop antenna current. From the circuit, it is seen that

$$I_2 / I_1 = (j\omega_o M) / (R_2 + j\,\omega_o M) \qquad (6\text{-}14)$$

If $R_2 << \omega_o M$, then I_2 / I_1 approaches unity, and this inequality need not be severe to be valid because complex numbers are involved.

The ratio $\omega_o M / R_2$ is that of the mutual loop antenna impedance to the loop antenna radiation resistance. It was shown to be

$$\omega_o M / R_2 = \frac{3}{4\pi} \quad \frac{\lambda}{r} \left[1 + \frac{\lambda_o}{j2\pi r} - \frac{\lambda_o^2}{4\pi^2 r^2} \right] \tag{6-15}$$

where r is the radial separation of the loops which lie in a plane.

The desired inequality is obtained for the condition $r << \lambda_o$, then

$$\omega_o M / R_2 = (3/16\pi^3)(\lambda_o / r)^3 \tag{6-16}$$

and the inequality grows rapidly as r becomes less than one-tenth of a wavelength.

Notice once more that the result is independent of loop size. The dimension of the loop vanishes from consideration because both $\omega_o M$ and R_2 vary as the square of the loop area; therefore the loop area cancels in the ratio.

Let us work a simple order of magnitude example to orient ourselves numerically. Suppose $r = \lambda_o / 100$, then $\omega_o M / R_2$ is of the order of 10^5. Then the magnitude of I_2 / I_1 becomes $10^5 / \sqrt{1 + 10^{10}} = 1$. Now suppose the loops are 0.1 meter in diameter and a frequency of 1000 Hz is employed. The wavelength is 3 x 10^5 meters and r = 3 x 10^3 meters. Thus, a 1 Ampere current in the 10 cm diameter transmitter loop will cause a 1 Ampere current in the tuned 10 cm diameter receiving loop located 3 kilometers away. It is difficult to visualize this behavior in the light of conventional experience. Of course, it would be even more difficult to reproduce this behavior experimentally because of the various loss mechanisms which introduce additional effective resistance in series with R_2.

If the condition $r << \lambda$ is approached by setting r and allowing λ to increase one can obtain this operation, in theory, for any frequency. At zero frequency, the implication is that a persistent current in a superconducting loop would set up a persistent current in another superconducting loop which did not originally

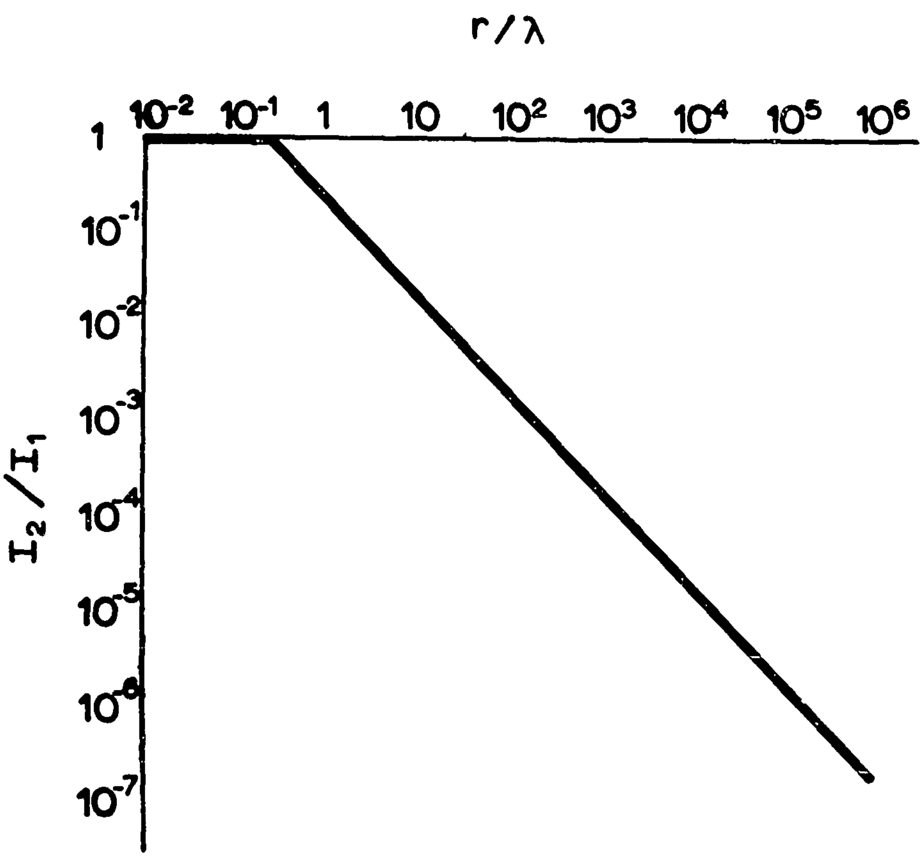

Figure 8. Illustrating the asymptotic behavior of receiving-transmitting current ratio for two ideal, identical, tuned, superconducting loop antennas of arbitrary size, spaced r/ λ normalized wavelength units.

carry a persistent current. Further, if the loops were of the same size, the two currents would be equal no matter how far the loops were separated. Such behavior is, of course, not observed. The theory can be saved by pleading an infinite build-up time because of the infinite Q.

The numerical example used above illustrates that "close range" does not necessarily imply short physical distances. When low frequencies are employed, this mode of operation could involve useful communication ranges. The superconducting loop would require many turns and the practical feasibility would hinge on the availability of extremely low-loss dielectric material for the tuning capacitor.

Figure 8 depicts the total performance for all modes of operation. Here the asymptotic behavior of receiving-transmitting current ratio is shown as a function of the loop to loop spacing in normalized wavelength units.

6.3 Loss Mechanisms

Substitution of numbers in the previous expressions leads to predictions that are somewhat spectacular in view of what is observed in actual practice. Nevertheless, there may be some element of reality in the theory because practical coils possess such large values of ohmic resistance that the necessary theoretical assumptions are quite invalid. Is it possible, however, to approach the predicted behavior if superconductivity is used to eliminate the ohmic resistance?

To answer this question it becomes necessary to identify and examine all the possible loss mechanisms which could prevent the realization of the theory. It is possible that some of the losses are avoidable through proper design and, in keeping with the objectives of this study, it is important to discover the extent to which superconductivity can assist in the reduction of all losses.

One more or less obvious loss mechanism is introduced by the process of detecting the receiving loop current. Any detector will have a maximum sensitivity, and this power requirement will show up as a direct or an effective series ohmic resistance in the loop. Dielectric losses in the insulating materials used to construct the loop and its tuning reactance will appear as effective series ohmic

resistance; this type of loss will appear whenever distributed material is within the near field of the loop. Concentrated parasitic structures near the loop will interact with the near field, by transformer action for example, and will introduce effective series ohmic resistance.

The alternating current behavior of superconductors will prevent full realization of zero ohmic resistance even though all other losses are eliminated. Finally, the effect of the propagation path must be taken into account. Thus, it is seen that superconductivity may furnish only a partial solution to the problem. The subsequent sections point the way to quantitative evaluation of the various losses and to a final assessment of the effectiveness of superconductivity.

6.4 Losses Due To High Frequency Effects In Superconductors

The theory of superconductivity predicts and experiment confirms a resistive loss due to alternating current operation of a superconductor [8, 40]. Arams, et al. [17] discuss this effect in connection with their work on a cryogenic radio frequency tuner. Their experimentally demonstrated Q shows, however, that at sub-microwave frequencies the loss is quite negligible in comparison to losses caused by other factors. In particular, the a.c. loss was seen to be orders of magnitude less than the dielectric loss encountered in the experiment. Since every effort was made to reduce the dielectric losses to a minimum, it can be concluded that the a.c. effects may be disregarded in any evaluation of a practical loop because other loss mechanisms will establish the minimum performance.

6.5 Dielectric Losses

The dielectric losses contemplated here are those due to the necessary supporting structures and insulating elements of the loop (or antenna, because these same considerations apply to any type of antenna structure) and its associated tuning reactance. In many cases the metallic conductors in the loop and the tuning capacitor can be rigid and nearly self supporting but even here some minimum quantity of dielectric material is required. At the lower frequencies where relatively many turns

of wire are needed and the tuning capacitance becomes large, it becomes impossible to avoid inter-turn insulation and solid capacitor dielectric materials.

The work of Arams, et al. [17] shows that dielectric losses can introduce a significant effective resistance in series with the radiation resistance. Radiation resistance was not involved in their experiment because the circuits were enclosed in a superconducting shield; however, the loss due to dielectric materials can be inferred from their data. It appears that a dielectric loss resistance as high as 10^{-3} ohm can be encountered.

If R_o is the effective ohmic resistance, a quick review of the derivation will show that

$$I_2 = [\omega M / (R_a + R_o)] I_1 \tag{6-17}$$

Where ωM is the mutual reactance and R_a is the radiation resistance. When the loss resistance is predominant, the expression may be written as follows:

$$I_2 \cong [\omega \mu_0 \beta^2 S^2) / (4 \pi r R_0)] I_1 \tag{6-18}$$

Where effective long range mutual inductance M has been replaced by its explicit formula. Here μ_0 is the absolute permeability, $\beta = 2\pi / \lambda_0$, and S is the cross-sectional area of the loop.

For a comparative example, choose a pair of 10 centimeter diameter tuned superconducting loops operating at 30 MHz and separated by 10^6 wavelengths. Let the radius of the wire in the receiving loop be 1 millimeter. Calculation gives the following values for the radiation resistance of a loop, the mutual reactance, the incuctive reactance of the receiving loop, and the idealized Q-factor of the receiving loop inductance:

$R_a = 1.9 \times 10^{-4}$ Ohm $\omega M = 4.8 \times 10^{-11}$ Ohm

$\omega L = 47$ Ohms $Q = \omega L / R_a = 2.5 \times 10^5$

Suppose that a 1 Ampere current flows in the transmitting loop. Then the idealized formula predicts a current of 0.25×10^{-6} Ampere in the receiving loop. Notice that this current is independent of the dimensions of the loops (although both loops should be of the same diameter).

The information bandwidth of the system is easily calculated. It is not impressive, being only 120 Hz: However, that is sufficient for some communication purposes.

If dielectric losses are present, the receiving loop current will be reduced. Moreover, the latter current no longer will be independent of the dimensions of the loop. This fact is evident from inspection of the formula above which accounts for these losses. In this particular example, the current is reduced by a factor of approximately 11 if the equivalent series resistance R_o of the dielectric losses were 10 times the radiation resistance, as it could be according to the referenced work.

The introduction of ohmic losses has several consequences. The first is the reduction of short circuit current, as was seen. The second is a decrease in the Q-factor. The third is an increase in signal bandwidth. The fourth is a decrease in the signal to noise ratio of the antenna. For the example discussed above, the Q is decreased by a factor of 11. The new bandwidth is approximately 11 times the former value.

The new signal to noise ratio requires careful interpretation here because the ohmic resistance will be at a very reduced temperature compared to the effective antenna temperature, in fact, the noise contribution from the cold ohmic resistance introduced by dielectric losses is negligible in comparison to that developed by the much smaller but very much warmer radiation resistance. In so far as signal to noise ratio is concerned, the deleterious effect of the ohmic resistance enters in different ways. First, an increase in signal bandwidth is an increase in noise bandwidth as well. Second, the available signal power is reduced by the addition of circuit resistance. In this particular instance, the new signal to noise ratio is approximately 9-percent of the ideal value. This is a remarkably small price to pay;

the reduction would be much more severe were it not for the fact that the cold ohmic resistance contributes little to the available noise power. The degree to which this advantage is retained depends upon how much the ratio (T_a / T_o), is greater than the ratio (R_o / R_a). Normally T_o will be greater than 300 degrees Kelvin. In fact, if the 30 MHz antenna is used for communications on the surface of the earth, T_a may well be in excess of 10, 000 degrees Kelvin. So long as the temperature ratio remains greater than ten times the resistance ratio, the rule of thumb is that the signal to noise ratio is in inverse proportion to the signal bandwidth ratio.

There is a relatively small penalty if the ohmic resistance is maintained at cryogenic temperatures. This fact may occasionally permit a trade-off of signal to noise ratio for increased signal bandwidth.

In conclusion, let us examine the performance of this communications system. The transmission path is 10^6 wavelengths or approximately 6000 miles. A realistic effective antenna temperature above the earth's surface is 10, 000 degrees Kelvin. If the transmitting loop current is increased to 10 amperes then the signal to noise ratio for the ideal system is approximately 20 (or 13 decibels) for the 120 Hz signal bandwidth. The signal to noise ratio for the system which includes the ohmic losses due to the supporting dielectrics is approximately 2 (or 3 decibels) for the 1200 Hz signal bandwidth. This remarkable performance (the transmitter requirements are less than 100 milliwatts) is still highly idealized because it does not take into account the near field environment of the antennas. And then again where on earth can one find a 6000 mile line of sight transmission path?

6.6 Losses Due To Nearby Parasitic Structures

The problem of the hot, near-field parasitic structure has been studied from the worst case point of view. The parasitic structure is represented by a tuned loop located at a distance r_{23} from the receiving loop. Its orientation is such that maximum coupling takes place. (Note that the radiation pattern of a dipole is such that unavoidable parasitic structures could be located judiciously in relatively "dead" areas. Here, however, we place the structure in the solid angle where it would have its greates effect.)

The analysis leads to the following expression:

$$\frac{I_2}{I_1} \approx \frac{\omega M_{21}}{R_2} \left[1 + \frac{\omega^2 M^2_{23}}{R_2 R_3} \right]^{-1} \tag{6-19}$$

where as before, the current ratio I_2 / I_1 is that of the receiving loop current to the transmitter loop current. Mutual reactance ωM_{21}, as before, is the far field mutual reactance between the transmitting and receiving loops. The resistance R_2 is, as before, the receiving loop radiation resistance plus the cold effective dielectric loss series resistance. The remainder of the expression shows the effect of the hot parasitic structure; obviously, it reduces the ideal current ratio. Mutual reactance ωM_{23} is that between the receiving loop and the parasitic loop and R_3 is the effective series resistance of the parasitic loop.

Calculations show that even for this worst case the reduction in current is only of the order of 0.1 percent. Further, since the position of the receiving loop generally is under control of the user, he can avoid the proximity of tuned elements or he can make sure that R_3 is large. It is anticipated that in most cases the presence of discrete, hot parasitic structures will not detract appreciably from the performance of the system.

Parasitic structures in the vicinity of the transmitting loop are of little concern; at worst, the effect would be an increase in transmitter power requirements.

Another interesting way of viewing the problem is to think of the parasitic structure in terms of an additional series resistance which is introduced into the receiving loop. Examination of the previous expression shows that this additional series equivalent resistance is given by

$$R_p = (\omega M_{23})^2 / R_3 \tag{6-20}$$

Possibly, the most serious concern is not the resistance itself but the fact that it is a hot resistance and will introduce additional noise power. The noise temperature

of R_p is the ambient temperature of the parasitic structure, or approximately 300 degrees Kelvin.

Relating the problem to our previous example system: if a high-Q, non-superconducting, 10-centimeter diameter, tuned loop were brought within 2 meters of the receiving loop, it would introduce an additional series resistance of the order of 10^{-8} Ohm.

Look at the problem this way. So long as the effective aperture of the parasitic structure does not seriously shadow the effective aperture of the receiving loop, there will be no significant reduction in performance. Thus, if the parasitic structure is electrically-small and non-superconducting, it will have an aperture of the order of the projected physical structure. The efficient electrically-small, superconducting loop will have an effective receiving aperture with a diameter that is approximately 0.4 wavelength. For our 30 MHz loop this receiving aperture would be a circular cross section 12 feet in diameter. Small screening metallic objects could easily be tolerated and fairly large dielectric objects could be placed in the effective area even though they were "semi-transparent" to the 30 MHz electromagnetic wave. Naturally there are faults with this particular point of view; it is suggested merely as a plausible intuitive aid in visualizing the problem.

6.7 Distributed Lossy Material In The Near Field

Ordinarily, something can be done about concentrated parasitic structures in the near field of the loop. Often, they can be removed, when recognized. If they are necessary, it may be possible to construct them of high resistivity materials and to avoid electromagnetic resonances near the loop frequency. Strategic orientation may place the structures in low field regions.

The problem is much more serious in the case of large quantities of bulk material or distributed material which is located in the near field of the loop. Sometimes the material will completely surround the loop, as for example the cryostat or Dewar containing the superconducting loop, or, again, the infinite half-space consisting of the earth located below the loop. When high frequency loops are employed, it may be possible to contain the near field in nearly lossless material; recall

that one-sixth wavelength represents the approximate radial distance where the near field and far field are equal in intensity. At low frequencies, however, where the wavelengths could be of the order of thousands of meters and more, it may be physically impossible to escape the near field.

Thus, the receiving loop (or any receiving antenna) may be positioned above the ground plane, say moist earth. Circuitwise, there is some coupling with this distributed material, currents are induced, and this coupling can be represented by an effective series resistance to be added to the other resistances present in the receiving loop. The situation is exactly that of the eddy currents induced in a transformer core by the alternating magnetic field of the winding. The resulting losses partly determine the value of the exciting branch resistance shown in the transformer equivalent circuit.

It is very important to recognize that the temperature of the effective series resistance due to the loss in the distributed material will be the same as the temperature of that material. The effective noise temperature of the total loop resistance will not be the ambient temperature of the loop but rather a temperature determined by the individual, different temperatures of the various resistance components which comprise the total loop resistance.

Again, we may need to account for the distributed lossy dielectrics and films in the structure of the cryostat that maintains the temperature of the superconductor at the proper value. Here, the temperatures of the effective series resistances will be low except for those caused by the outer layers of the cryostat.

Fortunately, a great deal of work has been done by previous investigators. Row has studied the magnetic dipole surrounded by a thin, lossy, spherical shell [41, 42]. We particularly call attention to his findings that it is possible to tolerate a much higher conductivity dielectric in the near field of a small magnetic dipole than with an electric dipole. This stems from the fact that the electric field intensity of the magnetic dipole varies at worst as the inverse square of the distance whereas the electric field of the electric dipole varies as the inverse cube in the near field.

Galejs has done very significant work in connection with the admittance of an insulated loop antenna in a dissipative medium [43]. He gives us exactly what we want, an expression for the effective series resistance introduced into the loop by the surrounding dissipative medium. If we call this resistance R_d, it can be found as a fraction of the loop radiation resistance by the following expression:

$$\frac{R_d}{R_a} \simeq \left[\frac{\sigma}{a f^2} \right] 10^6 \tag{6-21}$$

Where σ is the conductivity of the lossy medium in mhos per meter, a is the radial distance in meters of the lossy material from the loop, and f is the frequency in MHz. (Think of the loop as being in an empty spherical cavity, the walls of which are made of the lossy material.

For example, suppose the 30 MHz receiving loop were placed at the center of an underground spherical cave 60 feet in diameter and suppose the conductivity of the earth were 10^{-2} mho/meter. The additional series resistance introduced into the loop would be equal approximately to the radiation resistance of the loop. This additional resistance is an order of magnitude less than the previously investigated dielectric loss resistance due to the insulating material in the loop structure. The most important effect of R_d in this example would be a decrease in signal to noise ratio.

Most of the material in the cryostat surrounding the magnetic dipole will have a conductivity that is many orders of magnitude smaller than that of the above example so that the resistance introduced thereby will be much less than the radiation resistance, even though the material is three or four orders of magnitude closer to the loop. Over and above this, the laminar thickness of the thermally insulating layers will be much less than the skin depth for the electromagnetic wave; the expression, accordingly, represents an ultraconservative maximum. Finally, most of these resistances will be very cold, thus contributing little to the noise power.

It appears that the usual metallic coatings used for thermal reflection in Dewars must be avoided; however, this fact has not been completely evaluated at

this moment. Experience shows that such coatings are not absolutely necessary for practical cryostats.

Thus, the distributed material in the cryostat may be of minor consequence, but how about the ground plane or the equivalent of the ground plane? Bhattacharyya has studied the problem of input resistances of horizontal electric and vertical magnetic dipoles over a homogeneous ground [44], and has developed an expression from which the appropriate R_d may be calculated. It is rather complex in nature and the reader is referred to the reference for full details; however, Bhattachryya's graphs indicate that once the magnetic dipole is removed further than 0.3 wavelength (this is 9 feet for our 30 MHz loop) from the ground plane, the value of R_p falls to less than 10-percent of the radiation resistance of the loop.

It is interesting to note that considerably more resistance is introduced for the case of an electric dipole at the same height, thus agreeing with Row's assertion.

6.8 Transmission Losses

Something should be said about the propagation path between the transmitting loop and the receiving loop. Does the use of superconducting antennas modify, in any manner, the treatment of the space between the elements? If one excludes the immediate surroundings of the loops the answer is in the negative. At the site of the loop there is an advantage because the electrically-small superconducting magnetic dipole does not couple electrostatically through distributed capacitance to its region. But that is merely because of its small size, not because it is superconducting. Superconductivity enters only indirectly, since it makes practical the use of small elements.

In unbounded free space there would be no concern about the propagation path. The expressions which have been developed would give a true picture of long range magnetic coupling. However, if a dissipative medium lies between the two loops, the propagation losses are the same as for conventional antennas except at vanishingly small frequency.

A formal proof of the last assertion would go somewhat as follows. The attenuation of an electromagnetic wave occurs because the propagation constant,

$$\gamma = \sqrt{j\,\omega\,\mu\,(\sigma + j\,\omega\epsilon)} \tag{6-22}$$

possesses a real part. The most obvious source of a real part is the presence of a non zero value for σ, the conductivity of the medium. Not so obvious are possible contributions from the imaginary parts of μ, the permeability of the medium, and ϵ, the permittivity of the medium. The latter properties may be complex numbers. (As in plasmas, ferrites, and other interesting materials). However, let us simplify the problem by confining our attention to materials with real μ and ϵ.

The cause of the attenuation is the power dissipation in the medium and it is not difficult to trace power dissipation to a term in the energy balance expression, a term proportional to σE^2 where E is the electric field intensity. Consequently, (in our medium with real μ and ϵ) if no power dissipation is to take place it will be necessary that either σ is zero or E is zero.

In unbounded free space (no plasma) the conductivity is zero so that the presence of an electric field intensity can be tolerated. In a conducting medium, say in the earth or in sea water, the conductivity is non-zero so that an electric field intensity cannot be tolerated if propagation is to be attenuationless.

Now let us look at the electromagnetic field expressions appropriate to the loop. These are given by expressions (4-6, 4-7, 4-8) in section 4.2. Remembering that $\beta = 2\pi f/c$ it can be seen that an electric field intensity is present except when the frequency is zero. It is interesting to observe that the magnetic field intensity does not vanish at zero frequency; therefore, the long range magnetic coupling expressions would be valid for zero frequency no matter how great the conductivity. That this is true can be seen readily. The field of a permanent magnet penetrates a thick slab of copper without attenuation, provided there is no relative motion between the two items.

Unfortunately this limiting case is of no great value; the information

rate at vanishingly small frequencies is vanishingly small also. Other difficulties would be imposed if the medium contained significant quantities of unpaired magnetic dipole moments.

In conclusion then, the electromagnetic fields at useful frequencies are attenuated in the propagation path by a mechanism that is independent of the type of radiation which produces the fields.

6.9 Loading By Receiver Or Detector

The tuned receiving loop may be viewed either as a series resonant circuit or as an antiresonant circuit. In the former case one would wish to insert a zero resistance detector and in the latter case the detector, now connected in parallel with the loop and with the tuning capacitor, should possess an infinite input impedance. In either case no resistance, actual or effective, would be added to the total loop series resistance. In practice, both detector types represent limits to be approached; some resistance is added to the loop, its magnitude and its temperature are factors which will prevent realization of the theoretical performance.

It would appear at this time that detector loading need not be the primary loss factor. Dielectric losses and losses in distributed media may predominate. The role of the detector and quantitative estimates of detector performance are covered in some detail in chapter 7, which deals with methods of coupling to superconducting antennas. Since that material applies as well to the loop under consideration here we refer the reader to that source for further information.

6.10 Noise

A factor that is very pertinent to the long range coupling problem is noise, from the noise field present at the receiving antenna site. The discussion of thermal noise and induced "sky" noise in section 3 is completely applicable to the receiving loop in the coupled loop problem. The noise power spectrum will contain components which lie in the antenna bandwidth. These will excite the antennas and mask the receiving signal. When high-Q loops are used, the noise signal actually

has the appearance of a sinusoidal signal; this has a mechanical counterpart in the delicately suspended and isolated torsional pendulum which remains in constant detectable oscillation even though all intentional excitations have been removed.

It is unfortunate that the worst sky noise exists in the frequency range where the coupling under discussion would lead to the most interesting experiments. The expected magnitude of such noise must be calculated during the preliminary work on any such tests.

6.11 An Experimental Approach

It was discovered that Arams and his group [17] had already accomplished most of the early experimental objectives anticipated in this study. These objectives, as tabulated earlier in section 1.6, included experimental verification of high r.f. Q-values for superconducting tuned loops, high stability, and some verification of high coupling effectiveness. Since their work involved perfectly shielded circuits (A superconducting shield surrounded the tuned circuits), it remained to discover the effect on a superconducting tuned circuit of exposing the latter to the radiation environment before continuing with extensive experimentation.

The work has progressed to the stage where a tuned electrically-small, magnetic dipole has been operated in an r.f. transparent, liquid nitrogen Dewar. It is apparent that a number of technological problems face the investigator in this area. It is extremely important to assure that the experimental electrically-small antenna presents a perfectly balanced load at its terminals (with respect to ground); otherwise unbalanced r.f. currents will flow in lead wires and over attached equipment. When this happens, these latter items become part of the antenna system and may completely mask the performance of the antenna proper. The balance becomes more critical as the Q-factor of the tuned antenna increases. A second problem involves the detector. A true assessment of the performance of a receiving loop requires that the detector loss factor (the ohmic loss coupled into the antenna) be less than one or two times the antenna loss itself. Presently, high impedance types of detectors seem most suitable for initial work; even so, the best available balanced detectors leave something to be desired.

A third technological problem is the development of a sufficiently low loss r.f. transparent cryostat, or Dewar.

Our laboratory found that foamed polystyrene is eminently suitable for low loss r.f. cryostats in the liquid nitrogen temperature range. They are easily fabricated by primitive or by sophisticated means from blocks of the material, they are impervious to the liquid gas and will even hold a fairly good vacuum. It has been observed that tuned magnetic dipoles reach the Q-factor that is predicted when the temperature is reduced to 77 deg-K. This is in the r.f. transparent cryostat and the dipole may be immersed directly in the liquid nitrogen. The Q-factor increase is slightly more than 2.5 times the value at 300 deg-K. Operating frequencies were in the 15 MHz to 35 MHz range and the 77 deg-K Q-factors reached 1000 to 1500. The foamed polystyrene cryostats will operate for hours on a fill of liquid nitrogen; mission time would be no problem with this fluid.

Measurements of antenna Q-factor are important in these experiments. Ordinary Q-meters are useless for the high values encountered; consequently, it was necessary to develop a measurement technique which would give an accurate determination of the antenna Q without loading. A pulse excitation method was found to be most successful and most accurate and convenient. Two low-Q coupling loops are brought into the vicinity of the tuned magnetic dipole. One of them is attached to a pulse generator and the other to a high frequency, high gain cathode ray oscilloscope. If the magnetic dipole is removed, only the slightest coupling is observed between the two test loops. (One should take some pains to make certain that any natural resonances of the loops, leads and test equipment are reasonably well removed in frequency from that of the tuned magnetic dipole). Now when the tuned magnetic dipole is placed in position near the test loops and the exciting pulse is adjusted to the proper length, a perfect exponentially damped, train of waves at the natural resonant frequency of the magnetic dipole is observed. The coupling is made as loose as the sensitivity of the CRO and the output power of the pulse generator will allow. The Q-factor is easily and accurately calculated from the observed exponential decay by standard methods.

The theory indicated, with good certainty, that there exist other significant ohmic losses in the electrically-small antenna, such as will not be eliminated

by superconductivity. Our work still is limited by the lack of a suitable high input impedance detector and a means of making the latter physically compatible to the environment of the antenna. It appears that a miniature, balanced, high impedance, solid state preamplifier must be developed so that it can be placed in close proximity to the antenna before any realistic work can be done at superconducting temperatures. The added series antenna resistance caused by coupling leads to an external detector would balance out much of the hard-gained ohmic resistance reduction. Because the preamplifier must operate at near-zero temperatures, the ordinary solid state devices may not be suitable. A parametric amplifier seems a good possibility and the recently discussed indium arsenide, thin film, insulated gate transistor is reported to have a performance that is essentially constant down to and including liquid helium temperature [45].

Since conductor ohmic resistance is but one of the contributing factors to the total effective ohmic resistance of the antenna, the present idea is to use a series of identical antenna elements with successively higher Q-factors obtained by varying conductor materials and resorting to inexpensive cooling methods. The behavior of the superconducting antenna with its remaining ohmic losses could be obtained or predicted by extrapolating the curves showing the effects of increasing the Q-factor. In the superconducting antenna the residual ohmic losses are caused by unavoidable dielectric losses, coupling into parasitic structures, coupling into lossy distributed materials in the near field of the antenna (the earth, for example), and other factors that have been discussed earlier. These losses are great enough to limit the Q-factor of even the superconducting antenna to values that justify the asymptotic approach to the problem.

The most formidable problem, perhaps, is the cryostat or Dewar containing the superconducting antenna. There is no point in conducting experiments with a lossy Dewar, even though the tuned loop or antenna inside is in the superconducting phase. Conventional cryostats are out of the question because glass, aluminized films or any metallic heat reflecting films, metal bands, and the like will destroy the performance. A bold and imaginative approach is required. At present the investigation tends toward foamed polystyrene units using low r.f. dielectric loss plastics for mechanical reinforcement. Such cryostats work unusually well at liquid nitrogen temperatures. If a liquid nitrogen heat shield is used, it may be possible to extend the application to liquid helium temperatures.

SECTION VII

METHODS OF COUPLING TO SUPERCONDUCTING ANTENNAS

7.1 Introduction

In the case of transmitting antennas, the coupling problem usually involves efficiency. To what degree can one approach the theoretical efficiency of a particular mode of operation? For example, if maximum power transfer is the object, the theoretical efficiency would be 50-percent and it would be desirable to approach this value in practice. For an antenna, the useful power is that which is radiated, or circuitwise, that which is delivered to the radiation resistance.

When the antenna is electrically-small, the radiation resistance usually is less than the ohmic resistance, and it often is very much less than the ohmic resistance so that most of the power input to the antenna is wasted in uselessly heating this non-radiative portion of the antenna resistance. Superconductivity, which can eliminate the direct ohmic resistance of the conductor, promises to increase greatly the efficiency; however, the problem then shifts to the generator which now must have an internal resistance which is non-conventional in the light of existing technology. This difficulty was touched upon in section 1.7.

One solution is to use a matching section, also superconducting, so that the extremely small radiation resistance of the electrically-small antenna can be transformed to a more conventional value. Unfortunately, the result always includes bandwidth limitation because of the high Q-value of the electrically small antenna. While there are applications where the narrow bandwidth still is useful the solution, nevertheless, is limited to such special cases.

Another solution is to build the non-conventional transmitter with an internal resistance that is of the order of or lower than the radiation resistance. It

is not entirely impossible to do so, in fact, one technique can be suggested: The "transmitter" is a charged, low-loss capacitor and the modulator is a switch which can connect the charged capacitor to a superconducting electrically-small magnetic dipole. These elements would comprise a highly efficient, though relatively narrow band transmitting system. Refinements would require, among other things, superconducting leads and a superconducting, vacuum switch. Nor would there be any harm in placing the capacitor itself in the cryostat. Information would be transmitted by counting the number of cycles per burst (the switch must be opened only when the current passes through zero). Since this is a high-Q circuit, the system could operate for some time on an initial charge.

7.2 Ideal Detectors And Amplifiers

The circuit aspects of the antenna-detector interface have been discussed quantitatively in sections 4.3 and 4.4. Noise considerations are of primary importance. Relatively wideband operation is possible if the untuned electrically-small, superconducting dipole, electric or magnetic, is connected directly to a balanced infinite input resistance detector. (The use of the word resistance rather than impedance is deliberate). In fact, the bandwidth of the system would be determined by the amplifiers rather than by the antenna.

It was shown that the ideal detector or amplifier would be either of two types. The first type would possess an infinite input impedance, it would respond to the open circuit potential of the antenna and would, therefore, not draw a current from the source. The second type would possess a zero input impedance, it would respond to the current delivered by the source, the short circuit current of the antenna, and would, therefore, not require a source voltage.

The amplifier or detector-amplifier in the first stage of the receiving system should have considerable gain, voltage or current depending on the type, and it should be free of device noise. At least, the noise generated by the active devices should be less than some arbitrary value.

The electrically-small electric dipole is the natural source for the

infinite impedance type of detector or amplifier and the electrically-small magnetic dipole is the natural source for the zero impedance input. It is possible to invert this choice; however, in the one instance a magnetic dipole with the same effective length as a given electric dipole will turn out to be quite bulky in comparison, and in the other instance it would be difficult to justify a short electric dipole where one needs a source with a very low impedance. The inversion becomes less debatable when the input impedances of the amplifiers or detectors depart from the stated limits.

7.3 Practical Approaches To Ideal Shunt Mode Devices

We require an amplifier with an infinite input resistance and an infinite shunting reactance. Out of our everyday experience we would consider the modern cathode ray oscilloscope preamplifier as a practical approach. A typical 0-50 MHz CRO has an input impedance of 1 megohm in parallel with, say, 20 picofarads. With a low capacitance probe the input impedance goes to 10 megohms in parallel with, say, 5 picofarads. The first thing to be questioned is the input capacitance. It will form part of the tuning capacitance and must therefore be less than the required tuning capacitance, much less in fact because it does not have the quality of the main capacitor which will be made from superconducting material and which will have mainly helium for a dielectric. In the case of a 30 MHz magnetic dipole the antenna resonating capacitor typically is about 120 pf. The requirement seems satisfied.

Does the detector shunt capacitance have a sufficiently low loss factor? If the 5 pf probe is contemplated, its capacitive reactance at 30 MHz is about 1000 Ohms. If the loss tangent of the dielectric is 10^{-3}, the equivalent parallel resistance due to capacitor loss is 10^{6} Ohms; this is not good enough. Fortunately, there are practical dielectrics with loss factors better than 10^{-4}. The resulting equivalent parallel resistance would be 10^{7} Ohms.

It would be interesting to study the performance of this coupling method in the light of an application. Let us use the 30 MHz loop discussed in section 6.5 as a "typical" antenna. Here we are fairly sure that a practical loop can be built with a series resistance ($R_a + R_o$) of about 2×10^{-3} Ohm. We have seen that this would still be an excellent antenna in spite of the fact that the loss resistance R_o is about

ten times the radiation resistance. (The noise temperature of the loss resistance will be very low). The parallel equivalent resistance is given by the relation $R_p = Q^2 R_s$. Including R_o, the Q-value is about 2.5×10^4 in this case, thus $R_p = 1.2 \times 10^6$ Ohms.

The magnetic dipole antenna is a balanced system; consequently, it will be necessary to connect the output to a balanced detector unless a balun is employed. The latter alternative is a simple one for conventional antennas but it would create a variety of new problems here having to do, mainly, with the introduction of new losses even though the balun were superconducting. Let us specify a balanced detector because it is fairly easy to provide one.

Next, can the detector take advantage of the signal to noise ratio of the antenna? On the surface of the earth, and at 30 MHz a signal voltage which is just equal to the noise voltage at the parallel terminals is approximately 10 microvolts. If this seems large, remember that the actual antenna voltage V_a is multiplied in this case by the Q of the antenna circuit. Voltage V_a is only 0.4 nanovolt.

Further, the effective parallel resistance of the source is a complicated one in so far as noise generation is concerned, it is a composite of a hot resistance (R_a is at 10,000 deg-K) and a cold resistance (R_o is at 4.2 deg-K). One way to handle the calculation is to find the effective temperature of the 1.2 megohm resistance by the formula

$$\overline{V_n^2} = 4\,k\,TBR \qquad (7\text{-}1)$$

Substitution of numbers yields T = 1200 degrees Kelvin.

To summarize: We are dealing with a 1.2 megohm, 1200 deg-K source in a 1200 Hz band centered on 30 MHz. The source impedance reduces drastically on either side of the passband.

It has been shown in another study that the receiver noise figure which will just allow the detection of a signal that rises 11-percent (power wise) over the inherent source noise can be calculated from the following expression [46]

$$F_o = 1 + 0.26\,(T_a/T_o) \tag{7-2}$$

Where F_o = Required noise figure relative to T_o (as a ratio).
T_a = Effective antenna temperature, degrees Kelvin.
T_o = 290 degrees Kelvin (in normal situations).

Substitution yields $F_o = 2.1$ or, expressed in decibels $F_o = 3.2$ db. Conventional low noise preamplifiers are capable of even better performance at this frequency, accordingly we may conclude that practical detectors are available for utilizing the antenna's signal to noise ratio.

The final major problem is the actual connection between the antenna and the preamplifier. One might visualize a very short balanced line of high Z_o connecting the antenna to the balanced preamplifier which also is contained in the cryostat, but in a warmer location.

It would be interesting to compare, in performance, the non-superconducting, conventional, half-wave dipole. The 0.4 nanovolt signal can be related to the electric field intensity of the incident wave through the relation taken from page 323 of Schelkunoff and Friis [34]

$$V_a = E\,(2\pi/\lambda)\,S \tag{7-3}$$

Where S is the loop area of the one turn and E is the electric field intensity of the incident wave.

Substitution of our numbers yields an effective field intensity of 0.08 microvolt per meter. The V_a induced in the half wave dipole is found to be approximately 0.3 microvolt. The noise voltage due to the 73 ohm radiation resistance at 10,000 degrees Kelvin is approximately 0.2 microvolt for a 1200 Hz bandwidth. The conventional antenna is thus at least 3 db better than the electrically-small superconducting antenna, but this is to be expected in the light of earlier considerations.

For very low frequency antennas, there are interesting new devices which come much closer to the ideal infinite input resistance. A low noise, hybrid,

varactor-transistor device with a gain of 10^6 can operate at frequencies as high as 5000 Hz and can work from source resistances as high as 10^{12} Ohms [47].

The major difficulty revolves around the necessity for containing the preamplifier in the cryostat and in close proximity to the antenna. The amplifier must therefore be of minimum size and the active devices must be capable of operating at the low temperatures. Fortunately, none of these requirements appear to be insurmountable in the light of present technology.

7.4 Practical Approaches To Ideal Series Mode Devices

Low impedance detectors are not common. The previous 30 MHz antenna would require a series detector with an impedance below 2×10^{-3} Ohms, a seemingly impossible request. One might suggest a superconducting tuned radio frequency transformer to transform a high impedance to the low value but that only duplicates, with additional losses, what is done when the parallel equivalent circuit is used.

For radio frequencies below approximately 2 MHz there is a new type of amplifier that is amazingly compatible to the requirements. Newhouse and Edwards report an ultrasensitive linear cryotron amplifier [48] which has a vanishingly small series input resistance, mainly because the input control element is superconducting. While the input series resistance is at least three orders of magnitude below the maximum upper bound, there is some series input inductance. However, it is only of the order of a nanohenry and can be considered a part of the much larger loop antenna inductance. The minimum detectable signal is limited almost entirely by the radiation resistance temperature because the amplifier itself is a controlled superconductor and operates in liquid helium. The amplifier would be contained in the cryostat and its physical configuration is such that it could be in some cases an integral part of the electrically-small, superconducting antenna. The reference cited is most generous in detail and data; more excellent description is to be found in Newhouse's book [6].

Another most useful feature is the balanced arrangement. To minimize thermal fluctuations, the amplifiers are used in balanced pairs. The balanced input

is, of course, made to order for the loop antenna.

It seems safe to predict that the combination of a small superconducting tuned loop antenna and a linear cryotron amplifier could form the basis of a compact VLF (or ELF) receiving system with a sensitivity that is orders of magnitude better than what now is available. A system of this type should be unusually effective for underground transmission where the radiation resistance temperature is low, as was discussed previously. The losses in the transmission path cannot be avoided but additional range is made available by the increased signal to noise ratio.

SECTION VIII

MATERIALS AND STRUCTURAL TECHNOLOGY

8.1 Existing Superconducting Materials

A study of the considerable body of information on superconducting materials in the readily available literature will show that the existing types divide into two types. Type I superconductors are the so-called ideal or soft superconductors. These are the familiar materials known from the earlier days of the phenomenon. Type II superconductors comprise the non-ideal or hard superconductors; these are the new materials, the alloys, which have gained prominence through their use in high field superconducting electromagnets.

Lead would seem to be the most useful Type I superconductor for antenna construction because it is obtainable everywhere in sufficiently pure form, it is very inexpensive and easily fabricated into any shape by various non-critical techniques ranging from electroplating to casting. Its critical temperature of 7.18 deg-K is well above the temperature of liquid helium at atmospheric pressure and it can tolerate a reasonably high magnetic field. (The critical field is 803 Gauss). Lead is the logical choice for the low power superconducting receiving antenna. All the metallic parts of the antenna and its associated structure and circuit can be made from lead or lead coated metals.

Tin, with a critical temperature of 3.72 deg-K is a useful superconducting material for special applications. The temperature of liquid helium is reduced easily to this temperature by evaporative cooling at a lowered pressure. When tin is operated near the critical temperature it can be switched from the superconducting to the normal state by a small control current. This is made use of in the linear cryotron amplifier cited and described in section 7.4 and it could lead to a class of controlable, modulated, or variable bandwidth antennas.

Pure niobium is another interesting material because of its high critical temperature (9.46 deg-K) and high critical field (1944 Gauss), but it is expensive, more difficult to obtain, and does not lend itself to the fabrication of arbitrary shapes. Type I materials are not suitable for high power antennas, such as would be used for transmitting purposes, because the critical current which is related to the critical field through Silsbee's rule, is soon exceeded even at radiated powers of the order of one Watt, and less. This is true expecially for the electrically-small antennas where high-Q circuits lead to large circulating currents and where the extremely low radiation resistance requires a very large current for a given value of radiated power.

The Type II materials are tolerant of much greater fields and currents; however, according to the existing theory, these desirable properties are supposedly frequency limited to the extent that only d.c. applications are feasible. In fact, the earlier experimental evidence supported the theory. Now there is experimental evidence which indicates that high frequency, high power operation may be possible using these hard superconductors. Cummings and Wilson [22] find that there is a critical switching energy; short risetime pulses are supported until the power integrates to the switching energy before the Type II superconductor goes normal. Switching energies of approximately 12 Joules are reported for the Niobium-25-percent Zirconium alloy. The data cites operation of 15 mil superconductors at 300 Amperes and 15 kilovolts with risetimes of 2 nanoseconds.

8.2 Dielectric Materials

If it is important to locate conducting materials with infinite conductivity, it is equally important to find dielectric materials with near-zero dissipation factors. Tuning capacitors and supporting insulators are exposed to the alternating electric fields. It would be a source of frustration if the advantage gained through superconductivity were lost because of dielectric dissipation.

Our only lossless dielectric is a vacuum and some gases approach this quality closely. However, some mechanical supports are required, especially where stresses arise due to the extreme temperature variations to be encountered

in the case of superconducting devices. Again, thin film deposition may be an attractive technology for constructing miniature antennas; a low loss substrate would be necessary.

There appears to be a very limited choice of solid dielectrics. FEP TEFLON (Registered trademark of the E.I. DuPont Corp) is an outstanding dielectric, its properties improve as the temperature decreases. Allen and Nahman [19] give a particularly thorough theoretical analysis of the various loss mechanisms in this and other materials which may approach satisfactory performance at cryogenic temperatures. They report FEP TEFLON dielectric at liquid helium temperature as having a loss tangent of 7.3×10^{-6} with a possibility of 4×10^{-8}.

Foamed polystyrene would be a very useful low loss material; here the r.f. losses are reduced because of the low density of the medium. Quartz should be a suitable substrate material for the fabrication of deposited antennas, and D'Aiello and Prager [49] report that the Q-factor of rutile, which is of the order of several thousand at room temperature, approaches 10^5 at the liquid helium temperature.

8.3 Cryogenic Fluids

Liquid helium is practically the only cryogenic fluid that would be employed for the purpose of maintaining superconducting temperatures. Liquid hydrogen could be used for the superconductors with the highest transition temperatures; however, the risk is obvious and unnecessary. Other cryogenic fluids, liquid nitrogen primarily, are required for the secondary function of thermal shielding. The thermal properties, care and handling of cryogenic fluids are common information which does not need to be paraphrased here.

8.4 Operating Dewar Or Cryostat Requirements

The cryostat will be the major problem in an operating system utilizing a superconducting antenna. Other than supplying the low temperature environment for the antenna, the cryostat can but cause a deterioration in the performance. A superconducting antenna cannot tolerate lossy dielectric material, thin metal films,

metal structures and heat shields, and the like in its vicinity. The conventional Dewar will not serve, and even a Dewar with a geometrically extended radiation transparent region may not be satisfactory.

The analysis in section 6.7 shows that thin metallic heat reflecting films must be avoided and glass itself is not a suitable low loss dielectric for the radio frequencies of concern. "All-plastic" liquid helium Dewars are available on the commercial market, but they contain metallized plastic film, metal heat shields, and metal support rings and flanges. It is possible that such equipment still would function reasonably well as a Dewar if all the metal were eliminated. The liquid nitrogen shield and the evacuated walls should provide some operating life for a superconducting antenna.

Foamed polystyrene cryostats are very successful at liquid nitrogen temperatures. Many hours of mission time are available even in the case of crude units. The liquified gas is not observed to permeate the material at any appreciable rate. Since there is a gain in electrically-small antenna performance with any reduction in temperature (see section 4.7) this type of cryostat would give excellent service in that temperature range.

It is quite possible that an imaginative approach could extend the utility of this excellent r.f. media into the superconducting temperature range. The foamed polystyrene cryostat could provide a liquid nitrogen shield for an inserted liquid helium Dewar made from a low loss glass or plastic.

Finally, it must be realized that any gain in antenna miniaturization through superconductivity is offset by the bulk of the cryostat. An extremely large cryostat defeats the purpose even though it may be more efficient and have a greater mission time.

8.5 Prospects For New Superconducting Materials

Superconductivity is one of the most active areas of modern solid state physics and it is reasonable to expect new important discoveries, almost on a month to month basis. In particular, there seems some hope for new superconducting materials that will operate at much higher temperatures; the impact that such a

discovery would have on our subject is very apparent. The current status of the search has been outlined in section 1.11; it is seen that highly competent and responsible groups have taken the possibility seriously. But opinions are divided, and the reader is urged strongly, to study the Anderson and Matthias [23] discussion for an excellent background to the topic.

8.6 Possible Structures For Electrically-Small Superconducting Antennas

Several sketches of possible, electrically-small, superconducting antennas are gathered in a group at the end of this immediate section. Figure 9 shows the details of a 35 MHz magnetic dipole which has served as a trial design in our work. The coil has an inductance of 1.7 microhenry and the capacitance of the axial structure is 12 picofarads. The Q-factor for copper is in the neighborhood of 300. Lead is a suitable superconducting material here and the copper structure is lead plated after it is constructed. A standard lead acetate, sodium hydroxide, rosin solution gives satisfactory results. The plating bath temperature is about 170 deg-C. Electroplating voltage and current are 0.5 Ampere at 12 Volts.

It may be observed that the dielectric loss introduced by the TEFLON separators is minimized by increasing the capacitor spacing in those regions so that the capacitance at the ends is small compared to the capacitance of the central section. This is a very promising high-Q structure, the Q-factor has been observed to go to 1000 at liquid nitrogen temperatures for structures of this type. The center structure does not interfere with the behavior of the coil. The antenna can be fed by induction, or by shunt drive. The coil can be tapped symmetrically to provide a lower input impedance.

The technique used in the above antenna can be extended to frequencies ranging from 10 MHz to approximately 150 MHz. At frequencies lower than 10 MHz the number of turns increase to the point where a low loss coil form will be required. At extremely low frequencies it will be necessary to insulate the wire with a low loss dielectric. The tuning capacitor requirements at low and very low frequencies will necessitate the use of the best available dielectric material. It is conjectured that

Q-factors above 10^4 are available at the lowest useful frequencies.

The transmitting power capability of this type is limited if lead is used for the superconductor. High power operation may be possible under the conditions discussed in section 2 if Type II materials are used. Naturally, the mechanical design must accomodate the high electric fields that will be present in the capacitor and on various conductors; otherwise, ionization and arcing will occur.

Figure 10 shows a structure which could lead to a successful printed circuit antenna. Note that the lossy substrate, although exposed to the electric field, does not form the critical part of the dielectric. If the plated regions between the overlapping metallic areas are very closely spaced, the capacitor formed thereby will be very large in comparison to that due to the fringing field. Coupling can be accomplished by adjusting the relative sizes of the overlap areas; or, if loop coupling is desired, such a loop could be placed at an external location. Multiple dipoles could be printed on larger substrate areas. There are numerous variations to the shape of the coil. The actual material to be plated or deposited depends upon the power limit.

Another interesting form is shown in figure 11; these geometric shapes are printed on adjacent faces of two thin dielectric plates (as in figure 10), although it seems well worth testing thin low loss dielectric sheets with the patterns printed on the two sides. The spiral winding increases the inductance beyond that available with a single turn. The sectored Faraday shield increases the distributed capacitance between turns but still permits a radiating or absorbing dipole.

Figure 12 illustrates an interesting possibility. This tuned dipole, as was mentioned elsewhere, is really a combination of an electrically-small magnetic dipole and an electrically-small electric dipole. Both will radiate some of their stored energy but one or the other will be predominant, depending on its individual Q-factor and its effective electrical size; for example, the electrically-small antenna of figure 9 is predominantly, by far, a magnetic dipole.

In the unit illustrated here (figure 12) the capacitor structure is such that it becomes a more effective short electric dipole; in principle, it should be possible to make this element radiate as much power as the magnetic dipole which surrounds

it. We have, then, a unit which will radiate twice the power of the single element and it seems logical that the radiation resistance would double because the element current is unchanged. It is true that the two dipoles will be driven in phase quadrature because charge and current are in phase quadrature due to the way the stored energy shifts between the electric field and the magnetic field; however, this will not effect the modulation.

The radiation pattern is such that an extremely omnidirectional antenna results. The magnetic dipole produces a doughnut shaped pattern with the symmetry axis of the doughnut coincident with the axis of the coil. The doughnut shaped radiation pattern of the electric dipole has its symmetry axis perpendicular to that of the magnetic dipole; hence, it fills the gap in the former radiation pattern. Because of the phase quadrature of the driving signals and the space quadrature of the patterns, the pair of dipoles resemble a turnstile antenna at first glance; however, the E and H fields are interchanged in the electric and magnetic dipoles so that the two E fields (or the two H fields) are not in space quadrature as they are in the conventional turnstile antenna.

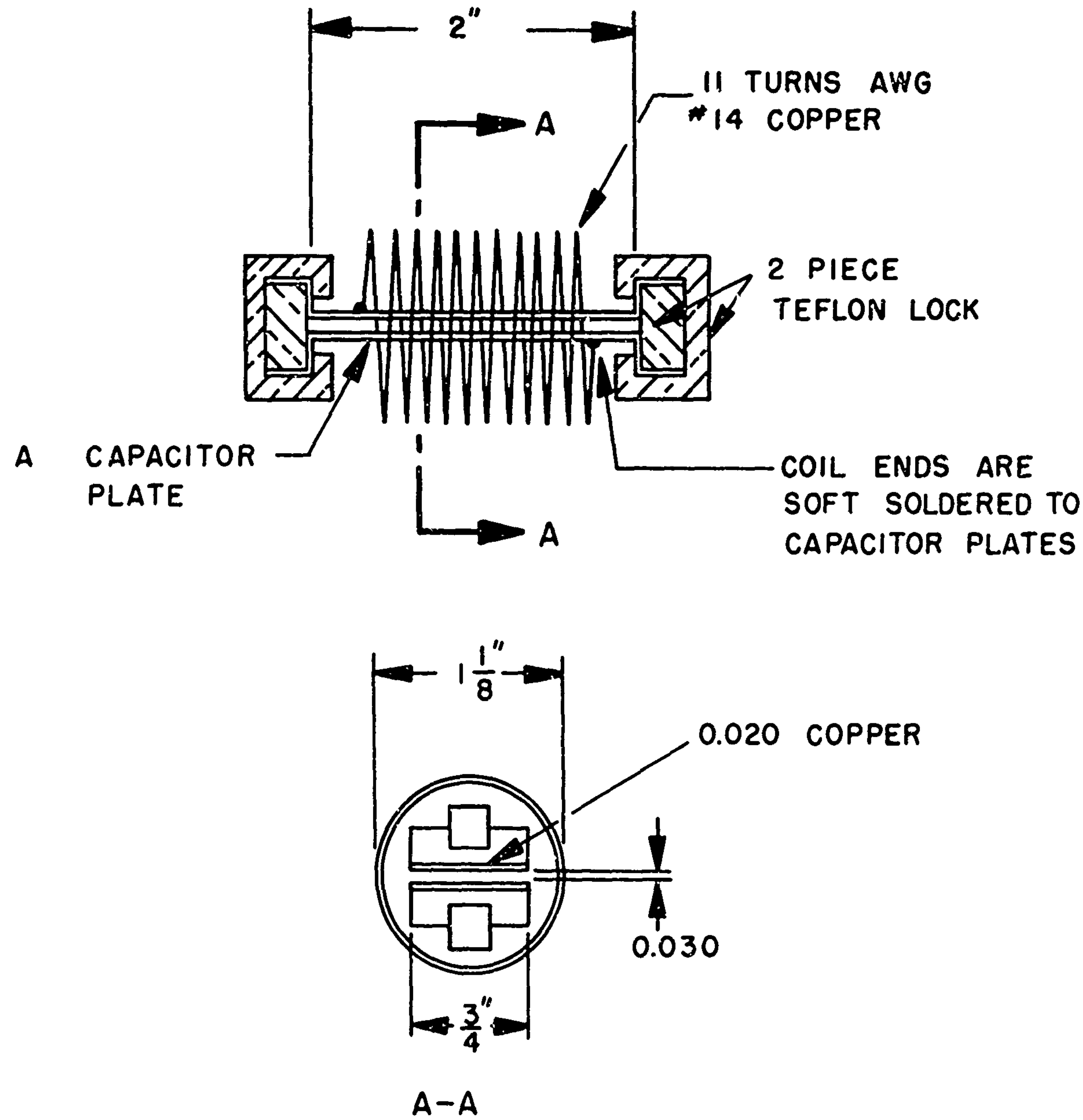

Figure 9. Dimensions Of A 35 MHz Tuned Magnetic Dipole.

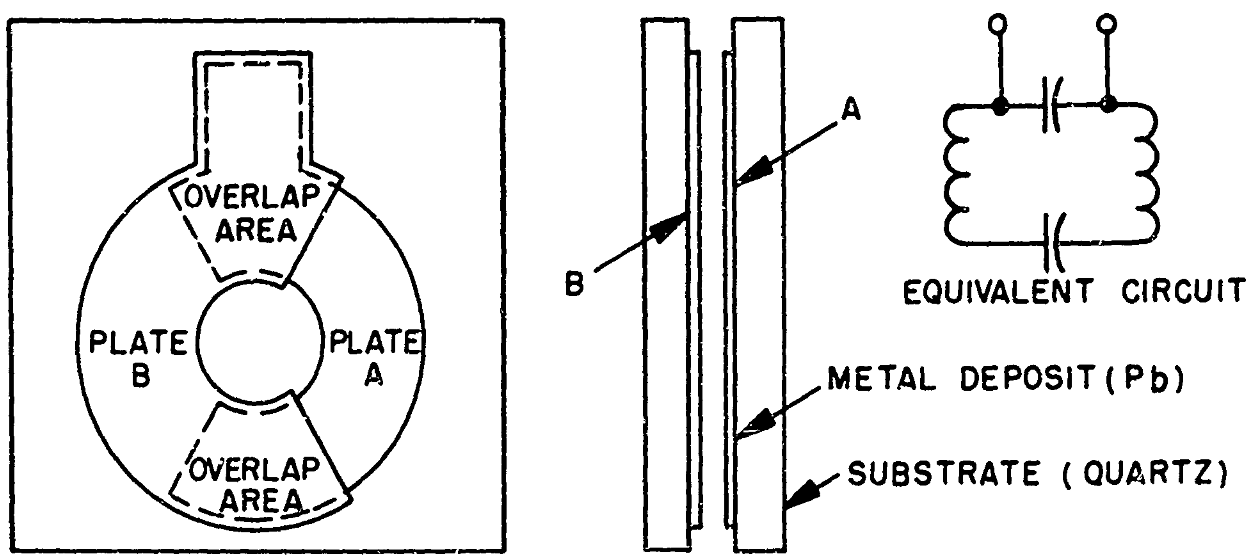

Figure 10. Scheme For Low Loss Magnetic Dipole

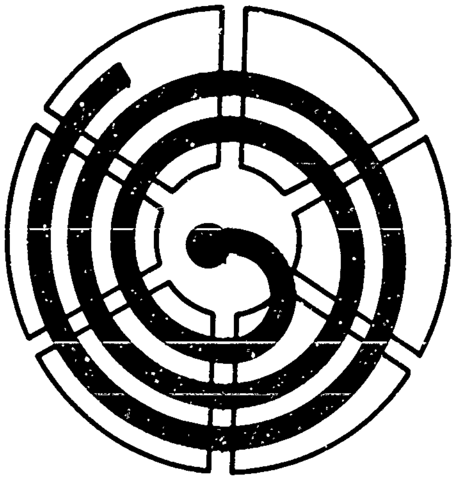

Figure 11. A Possible Printed Antenna.

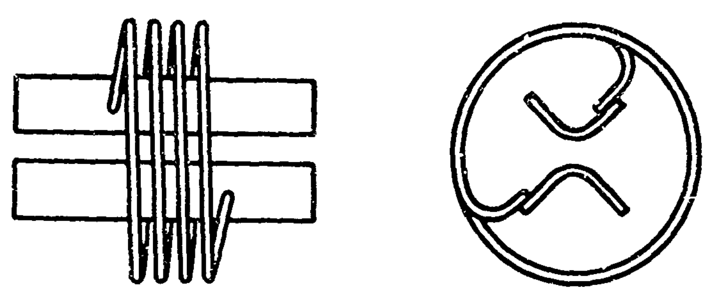

Figure 12. A Miniature Dual-Dipole Antenna.

SECTION IX

QUANTUM EFFECT ANTENNAS

9.1 Quantum Effects In Superconductors

There are various quantum effects in superconductors, some of which may lead to antenna-like applications. These are the non-conventional applications of superconductivity discussed in the opening section. It is possible that innovative antennas may result from such application because the mechanism of interception is different. Greater bandwidth is one such possibility because electric circuit parameters such as inductance and capacitance are not directly involved. A somewhat broader interpretation of the word "antenna" may be needed in some cases. If one accepts the definition, "An antenna is a transducer between the electric circuit and free space, al' ing energy (usually modulated by information) to be transferred from the one to the other," then it is possible to cite existing experimental evidence from the literature demonstrating both the the function of transmission and that of reception. At the present moment the prospect for receiving antennas is greater.

The popular literature contains numerous well written accounts of the quantum effects and the reader is urged to survey these for additional background. Bloch [12] and Parks [9.10] discuss some of the important quantum effects dealing with magnetic behavior. The Josephson effects are under intensive current investigation by many laboratories; papers and reports appear almost weekly in numerous professional and scientific journals; Langenberg, et al. [11] give a readable account and the current abstracts will lead to many other articles.

Quantum effects in superconductors and quantum effects that may be related to superconducting, electrically-small antennas enter our specific area of concern from various directions. They enter, for example, when one searches for limiting conditions on the minimum physical size of an antenna for a given frequency.

Consider the electrically-small, resonant loop antenna that has been studied by our group. In its simplest form it consists of a circular ring that is broken for series insertion of a parallel plate capacitor. All materials including the capacitor plates, are ideal (soft) superconductors. What we have is a simple resonator. The loop is the inductance, the capacitor resonates the circuit at angular frequency $(LC)^{-\frac{1}{2}}$.

Notice that this is not a closed ring. It will not support a persistent current and trap flux in the manner of a closed ring. It is an extremely high-Q circuit and will store electromagnetic energy.

Now let us pose a question. Suppose the dimensions of the ring and capacitor are reduced, and we are particularly interested in reducing the spacing between the plates of the capacitor. Classical electromagnetic theory will continue to predict a resonant frequency in accordance with the previously stated expression; however, what will happen when the spacing between the plates is reduced to an order of magnitude which supports a Josephson effect?

There are microscopic quantum effects which allow radio frequency waves (with no lower frequency limit) to be radiated or absorbed by objects as small as an individual atom. These are not necessarily related to superconductivity, but there is sufficient analogy to raise occasional comment in the literature. These usually amount to an observation that the superconductor is a macroscopic quantum state and that it exhibits quantum phenomena which parallel the quantum behavior of microscopic systems. Should we not, therefore, turn to superconductivity in our search for a macroscopic duplication of the microscopic quantum effects active in radiative transfer between the atom and the radiation field?

9.2 Radiation From Persistent Currents

It is sometimes suggested in the literature that superconductivity is a manifestation of quantum mechanics operating on a macroscopic scale, that a piece of metal in a superconducting state is somewhat like a single giant atom. In this light, it becomes interesting to carry out an exercise which parallels the classical prediction of radiation by an orbital electron because of its constant acceleration

in the circular orbit. (Such radiation is not observed and, of course, such a model would predict a catastrophic end for the atom. The quantum theory has replaced this model with the concept of stable, metastable, and temporary non-radiating states). Suppose that a closed superconducting ring carrying a persistent current is like the current of the electron orbit model, what is the outcome of a calculation?

Thus, according to the classical theory, accelerated charges radiate; if this were so for the superconducting ring, which contains accelerated charge, the persistent current should decay because of the loss in energy.

The basis of analysis is the classical expression

$$dW/dt = (2/3)\ (e^2/c^3)\ \underline{\dot{r}}^{\cdot\,2} \qquad (9\text{-}1)$$

Where dW/dt is the rate of energy loss, e is the electronic charge, c is the speed of light, and $\underline{\dot{r}}^{\cdot}$ is the acceleration vector.

For the charged particle in a circular orbit (we are thinking of the free electron in the superconducting ring) in the X-Y plane, the particle displacement is given by

$$x = r \cos \omega_o t \qquad (9\text{-}2)$$

$$y = r \sin \omega_o t \qquad (9\text{-}3)$$

Where r is the radius, $\omega_o = 2\pi f_o$ where f_o is the frequency of rotation; also $\omega_o = v/r$ where v is the tangential speed of the charge.

After two time derivatives

$$\ddot{x} = -r\,\omega_o^2 \cos \omega_o t \qquad (9\text{-}4)$$

$$\ddot{y} = -r\,\omega_o^2 \sin \omega_o t \qquad (9\text{-}5)$$

and since

$$\left|\ddot{\underline{r}}\right|^2 = \ddot{x}^2 + \ddot{y}^2 \tag{9-6}$$

then

$$\left|\ddot{\underline{r}}\right|^2 = r^2 \omega_o^4 \tag{9-7}$$

In other words, charges moving in a circular path are constantly accelerated. Finally

$$dW/dt = (2/3)\ (e^2/c^3)\ r^2\ \omega_o^4 \text{ ergs/second} \tag{9-8}$$

If $\underline{v}$ is the drift average velocity associated with a density n electrons per unit volume and if e is the charge per electron, then the current density associated with the charge motion is

$$\underline{J} = ne\ \underline{v} \tag{9-9}$$

and for circular motion

$$\omega_o = v/r = J/ner \tag{9-10}$$

then

$$dW/dt = (2/3)\ J^4/(c^3\ e^2\ r^2\ n^4) \text{ ergs/sec} \tag{9-11}$$

where

J is in esu per sq cm

e is in esu

n is in electrons per cubic centimeter

r is in centimeters

If the accelerated free electrons in the ring do radiate electromagnetic energy, as they should, the radiated power and frequency can be determined by the above results. Suppose we deal with a very small ring of lead, only 10 microns in radius. (Persistent currents have been observed in rings of this size [10].)

There is a limit to the supercurrent density J. This is given to within an order of magnitude by:

$$H_c = J_c / \lambda \tag{9-12}$$

Where H_c is the critical magnetic field, J_c is the corresponding critical (maximum) supercurrent density, and λ is the penetration depth, see Newhouse [6] chapter 3.

For lead $\lambda \cong 10^{-5}$ cm and $H_c \cong 800$ Oersteds. This yields a J of nearly 10^{-8} amperes per square centimeter in the surface layer of the superconductor.

When values are substituted, ω_0 is of the order of magnitude of 10^7 radians per second. This corresponds to a radiation frequency near 10^6 Hz.

Further substitution in the energy expression yields a radiated power of 10^{-28} watt, which in turn can be written as -250 dbm.

The inductance of the small ring will be in the neighborhood of 30 picohenrys and the actual current in the ring calculates to be about 1 Ampere. The stored magnetic energy of the ring from $W = (1/2)\ LI^2$ is approximately 10^{-11} Watt-seconds.

In a purely brute force way we can combine the two results (10^{-11} Watt-seconds)/(10^{-28}Watt) = 10^{17} seconds.

There are about 3×10^9 seconds in a century; accordingly, it would require a considerable number of centuries for the appearance of even the slightest indication of a current decay.

Thus, all this work is inconclusive. If the ring is radiating a 1 MHz signal, we have no device sensitive enough to detect the radiation, -250 dbm sensitivity is well beyond any present technology. Further, we cannot tell if the

persistent current is decaying because measurement of the predicted decay is impossible due to its slowness.

To summarize, if the persistent current in a superconducting ring can be likened to a non-radiating Bohr orbit, or a stable quantum state, the classical theory at least does not predict an energy loss rate that could contradict this hypothesis by a simple measurement. Radiation is predicted, but it is far too small to be measured by any present or future device, therefore, we cannot prove the validity of the classical theory either, although experiments show it to be true for high energy charged particles, especially electrons, moving in a circular orbit. One is safe in assuming a stable quantum state for the persistent current in the ring if he wishes to do so.

9.3 Radiative Transitions Between Persistent Current States

Pursuing the analogy further, the next step is to compare the persistent current in a closed superconducting ring to the non-radiating state or "orbit" of the atom. In the atom, there are radiative transitions between non-radiating states.

Given that the superconducting ring represents a stable, non-radiating quantum state, it would seem plausible that changes from one persistent current to another, if such changes could be produced without forcing the ring to go normal, could be interpreted as behavior equivalent to the radiative transitions between energy levels of an atom.

It is known that the persistent current in a closed superconducting ring is quantized via the quantized flux. The theory is well confirmed by experimental evidence [9], [10], [12].

The size of the flux quantum is :

$$\Delta \Phi = \frac{h}{2e} \text{ MKS Units (Webers)} \qquad (9\text{-}13)$$

where h = Planck's constant and e is the electronic charge. It would be interesting to examine the amount of energy involved when a superconducting ring undergoes a

transition from one quantum state to another. If the amount of energy corresponds to that of an electromagnetic photon in the radio frequency range, then one could foresee an antenna-like function where the ring might absorb such quanta from an incident electromagnetic wave; or, where transitions from a higher to a lower energy state could perhaps be associated with radiation of electromagnetic photons from the ring.

The energy state of the ring can be found from:

$$W = (1/2)\ LI^2 \tag{9-14}$$

where L is the inductance and I is the persistent current. This is the familiar expression for the energy stored by a current carrying inductor.

Inductance for a 1-turn loop is defined by:

$$L = \Phi/I = d\Phi/dI \tag{9-15}$$

(the later being true because L is a constant for a substance with constant permeability) hence:

$$I\ d\Phi = \Phi\ dI \tag{9-16}$$

Substitution for L in the energy expression yields:

$$W = (1/2)\ \Phi I \tag{9-17}$$

and differentiation gives:

$$dW = 1/2\ \Phi\ dI + 1/2\ I d\Phi \tag{9-18}$$

Replacement of the first term on the right by a previous equality results in

$$dW = I\ d\Phi \tag{9-19}$$

But the minimum $d\Phi$ is the flux quantum $h/2e$; therefore the energy quantum corresponding to a transition to an adjacent energy state is:

$$W_1 - W_2 = I\,(h/2e) \tag{9-20}$$

Where the numerical subscripts denote the stable energies of two adjacent states.

If an electromagnetic quantum is radiated or absorbed, it must be true that:

$$W_1 - W_2 = hf \tag{9-21}$$

Where h is Planck's constant and f is the frequency of the radiation. Comparing the last two expressions we find that:

$$f = I/2e \tag{9-22}$$

predicts the frequency of the radiated or absorbed photon.

To test this relation, use $I = 1$ Ampere as a typical value for the persistent current. The MKS system value for e is 1.6×10^{-19} coulomb. We calculate:

$$f = 0.31 \times 10^{19} \text{ Hz} \tag{9-23}$$

which seems a somewhat ridiculous value because this has the frequency of a hard X-ray photon.

At around 150 microamperes for the persistent current the hypothetical transition energies correspond to visible light; and at 1 nanoampere we are talking about microwave frequencies around 3100 MHz.

All this would have to be examined in the light of the superconductor energy gap theory [7], [50], which explains why superconductivity disappears at

extremely high frequencies. For normal radio frequencies at least, the hypothetical behavior would be an increase in the persistent current upon illumination of the ring by radio frequency energy of the proper frequency.

One reason why a phenomenon of this type could have escaped detection is the extremely small persistent currents corresponding to frequencies of interest; persistent currents induced originally by a field as weak as the earth's field still are too large. Very careful shielding would be required to develop the proper "bias" field.

A second reason would be the lack of a suitable detector. Now, however, by using the Superconducting Quantum Interference Device, which is to be described in the next section, in conjunction with an integrating digital counter, it would be a simple matter to observe changes in flux brought about by incident electromagnetic radiation. This latter behavior would be equivalent to that of an antenna with a sensitivity beyond anything now available. This hypothesis seems interesting enough to warrant some further investigation.

9.4 Realization of A Quantum Effect Antenna

The recently announced superconducting quantum interference device [51, 52], the SQUID, developed by the Ford Scientific Laboratory, Dearborn, Michigan, uses a quantum effect to measure weak magnetic fields. While it has not been so stated, there is no reason why the device cannot be considered a receiving antenna. Forgacs and Warnick [52] already have used it to measure the micropulsations present in the vertical component of the earth's magnetic field. In this application the sensitivity of the SQUID compares very favorably with, and may well exceed, that of the 2 meter diameter, 21,586 turn loop antenna used by Campbell [53] to measure the same fields.

It is well known that a closed superconducting ring traps magnetic flux and further that the trapped flux levels are quantized [9], [10], [12]. The device under consideration here, by forcing the operating point of a superconducting ring to hover in the transition region between the superconducting and the normal (non-superconducting) states, allows the ring to make changes in the amount of trapped flux.

These changes must, however, occur in quantized (or digitized) steps and the size of each step is exactly known. The steps can be counted and an interpolation between steps is possible because of the peculiar nature of the phenomenon.

The flux contained in the ring may be likened to a jar containing a known number of identical balls. If one keeps track of the number of balls taken out or added, the net number in the jar is known at any time. Likewise, the "trapped" flux, which continues to adjust itself in accordance with the magnetic field in the region where the ring exists, is known exactly because the identical incremental changes are counted.

The electronic system which permits the flux density determination is a carrier feedback system. The size of the error signal is in a one-to-one correspondence with the flux density. The flux quanta, or "increments" are so small that the process permits extremely high resolution, and, as was previously mentioned, there is a method of interpolation which quantifies the fractional part of an increment.

Physically, the device consists of a tiny probe (The SQUID element itself is about 1/2 inch in diameter, representing quite a contrast in size to Campbells antenna), plus the electronics and the Dewar. The electronics are not excessive and could be miniaturized; the Dewar probably would be the largest element in the system.

The particular system reported in the literature uses a 10 kHz carrier and has a system bandwidth of 4 Hz. The device itself does not determine the bandwidth, it is not to be compared with any type of tuned circuit or conventional antenna. Higher carrier frequencies and greater system bandwidths should be feasible if the purpose of the system were to be directed toward an antenna application. At present the intended application of the device is that of measuring extremely weak magnetic fields. Field strengths below 10^{-8} Gauss can be measured accurately. In particular, the minimum detectable field change (or signal) is reported to be between 3×10^{-9} and 4×10^{-8} Gauss. In a plane wave this would correspond to an electric field intensity of approximately 100 microvolts per meter,. By way of comparison, Campbell's system used a 1 Hz bandwidth and worked in a noise level of 10^{-7} Gauss.

SECTION X

PROSPECTS FOR ELECTRICALLY-SMALL, SUPERCONDUCTING ANTENNAS

10.1 Summary: Advantages And Limitations Of Superconducting Antennas

The advantages of superconducting, electrically-small antennas are not so clear-cut that a neat and unqualified summary can be formulated. Some of the limitations are rather evident and often the advantages seem marginal in view of the technological difficulties and inconveniences generated by the need for temperatures near absolute zero. There is no manifest behavior of a superconducting antenna which results in extraordinary or unusual performance, compared to the conventional antenna. There are advantages, but often they are qualified in nature and always they are hard won. Sometimes they are marginal.

The following is a collection of statements which have been found applicable to electrically-small antennas; they carry reasonable assurance and are offered as a summary:

a. The possibility of miniaturization represents the principal advantage of a superconducting antenna; most other advantages follow as a result of this possibility.

b. The superconducting, electrically-small transmitting antenna has a much greater radiation efficiency than a normal antenna of the same size; however, there is a corresponding bandwidth reduction which may make the efficient antenna impractical for some applications.

c. The superconducting, electrically-small antenna (transmitting or receiving) cannot compete in electrical performance with a non-superconducting half-wave dipole but can only approach the

performance of the latter in the limit. The one exception may be in the case of a receiving antenna used with a perfect receiver.

d. For a given bandwidth the superconducting electrically-small receiving antenna can have a greater antenna signal to noise ratio than a non-superconducting, electrically-small antenna of comparable size. This advantage disappears as the bulk of the cryostat containing the superconducting antenna becomes large in comparison with the conventional electrically-small antenna and any associated bandwidth reduction elements or preselectors.

e. In many and perhaps most instances, superconductivity does not eliminate all effective ohmic resistance in the antenna circuit; consequently, it is debatable whether or not a truly superconducting antenna can exist. (The word, "superconducting", is taken in its narrow sense here).

f. The superconducting, electrically-small receiving antenna provides an increase in signal to noise ratio at the expense of bandwidth; however, the gain in signal to noise ratio greatly exceeds the loss in bandwidth.

g. An electrically-small receiving antenna need not utilize superconductivity in order to achieve a significant gain in signal to noise ratio. Such performance gain accompanies any reduction in temperature and a very important improvement takes place at liquid nitrogen temperatures. Actually, in most instances it is not entirely clear as to whether the greatest benefit is gained by the extreme temperature reduction which is necessary to produce superconductivity or by the elimination of the conductor ohmic resistance.

h. The greatly improved signal to noise ratio of a superconducting, electrically-small receiving antenna is available for broadband operation to a high-impedance, low noise receiver.

i. The superconducting, electrically-small antenna represents a practical approach to a point source dipole radiator (or absorber) and would be useful in close spaced arrays where the physical size of normal antenna elements leads to unwanted coupling.

j. The superconducting, electrically-small receiving antenna offers a means of achieving an extremely narrow bandwidth antenna with excellent sensitivity and stability. It is far superior in these respects to "Q-multiplying" circuits utilizing active elements.

k. Electrically-small devices utilizing quantum effects in superconductors can be employed as receiving antennas. Here, superconductivity enters in a radically different way and the device can be much smaller than the electrically-small antenna using superconductivity in the conventional manner. Further, the bandwidth limitation does not apply to the quantum effect antenna.

At this moment it is not possible to offer a high incentive for replacing conventional large antennas by electrically-small, superconducting antennas. There are, however, possible instances where it would be useful to substitute an electrically-small superconducting antenna or possibly a cold, non-superconducting, electrically-small antenna for a warm, non-superconducting, electrically-small antenna. Whenever and wherever an antenna of the latter type is being used and only marginal performance is being obtained, it becomes worthwhile to consider the substitution.

10.2 Special Purpose Applications For Superconducting Antennas

It seems clear that superconducting antennas will not find general application because of the low temperature requirement, because of the narrow bandwidth, and because they seem to offer mainly the possibility of miniaturization. Within these limitations lie a number of special applications that are worth enumeration and consideration. We list a few examples, in no particular order, to illustrate the point:

a. Extremely small but still useful dipoles are possible. These represent what might be called point source dipole radiators and absorbers. Greater close range precision is possible when these elements are used to drive precision reflectors or when they are formed into precision arrays.

b. An extremely high-Q tuned antenna will ring for an appreciable period of time when it is pulse excited. In this way it can function as a storage device, holding some characteristic of an incoming signal. An array of such antennas all tuned to different frequencies and spaced sufficiently far for negligible interaction could store a sampled spectrum.

c. The extremely narrow bandwidth antenna still is useful in transmitting or receiving a single frequency used as a standard or used as part of an electromagnetic, spatial reference grid.

d. The extremely high-Q magnetic dipole has a large electromagnetic cross section and is sensitive to objects in its near field. A superconducting tuned dipole is inherently stable and sensitive and should outperform anomaly locating instruments employing active devices and using a feedback mode for generating high Q-factors.

e. In the LF, VLF, and ELF ranges where even conventional antennas of reasonable physical size are necessarily electrically-small, the superconducting (or even cold) electrically-small magnetic dipole offers a gain in signal to noise ratio and/or further reduction in physical size. Such antennas may be useful in measuring sky noise or galactic noise from vehicles that cannot support large structures. In space platforms such antennas could be placed above ionized regions which absorb or modify low frequency radiation and it may be possible to utilize the natural cooling which is potentially available in space. It will be recalled that the receiving antenna need not be superconducting to achieve important gains in signal to noise

ratio. Also, natural cooling would be more feasible with a small structure than a large structure because of the necessity for a radiation shield. If directivity were required, two properly spaced dipoles could form an interferometer-type array. The narrow bandwidth of the high-Q antennas might be an asset because the measured frequency is well defined.

10.3 Directions For Further Research

There are several areas where further research would be of value. High on the list is the need for experimental investigation of the high power pulsed r.f. behavior of the Type II superconducting coaxial transmission line as used by Cummings and Wilson [22]. Any success here would lead immediately to a realization of high power, superconducting, electrically-small transmitting antennas.

Superconducting antennas will not become practicable unless a suitable r.f. transparent Dewar or cryostat is feasible, and the word "suitable" carries many implications. While this is a matter of technology, it remains, nevertheless, a matter of vital importance.

It is possible on theoretical grounds to justify experimental research in deep strata communications using very low frequencies and superconducting, electrically-small magnetic dipoles in association with innovative receivers utilizing cryogenic amplifiers or certain modern solid state amplifiers that are operated at cryogenic temperatures.

Section 10.2 suggests several special applications for the electrically-small, superconducting antennas; one or the other of these would furnish the basis for an interesting research project. Some other topics mentioned in the report warrant further investigation: Natural cooling and the behavior of a self resonant, superconducting coil (as a possible wide bandwidth, electrically-small, superconducting magnetic dipole) are two such items.

The application to antennas of certain quantum effects in superconductors seems to be a fertile field for further research, and there is some justification

for a new look at superdirective arrays.

This report would be incomplete without the last minute inclusion of a citation of new experimental results just disclosed in the literature by Siegel, Domchick, and Arams [54]. These investigators have operated a self resonant coil made of a Type II superconductor (Nb_3Sn) at 26.8 MHz, achieving an unloaded Q of 3.7×10^6 at 4.2 deg-K and 0.45×10^6 at 16.2 deg-K. The circuit was observed to have a relatively high power handling capability. Their results have an important bearing on the discussion of high frequency behavior of Type II superconductors, section 1.5; the conjecture regarding self resonant coils, section 5.1; and the optimism regarding maximum radio frequency signal levels, section 2.4.

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13 ABSTRACT The advantages and limitations of electrically-small, superconducting antennas have been investigated. The study led to a consideration of miniaturization, physical shape factors, long range magnetic coupling, maximum signal levels, antenna-receiver interface problems, materials, structures, and potential antenna applications of the quantum effects in superconductors. In addition, natural cooling and superdirectivity were incidental but relevant topics. In general, it was found that the possibility for miniaturization represents the principal advantage of the superconducting antenna, especially at the lower frequencies where antennas often are electrically-small through physical necessity. Radiation efficiency is increased in transmitting antennas, but at the expense of bandwidth. The degree of usefulness of superconductivity in receiving antennas depends considerably on the low noise properties and input impedance of the receiver and on the environment of the antenna. Any cooling improves the performance of the antenna; in fact, except in the case of the perfect receiver, there is some question as to which is more effective, the low temperature or the superconducting mode.

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